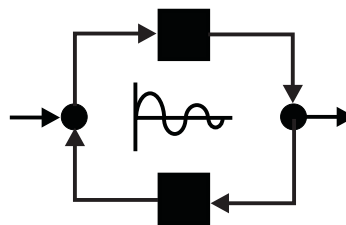


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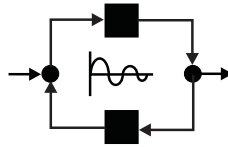
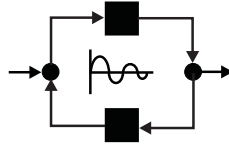


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Forward and Inverse Fuzzy Magnetorheological Damper Models for Control Purposes

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Abstract

The magnetorheological (MR) damper behavior is described by a set of nonlinear differential equations, which are unsuitable in control applications. This paper illustrates the effectiveness of adaptive neuro-fuzzy inference system (ANFIS) in building both forward and inverse MR damper models. The forward model is obtained by training noisy data generated from simulation of the nonlinear differential equations with low pass filter. Input selection method is applied to find the most appropriate inputs to train the inverse model. Finally, the numerical simulation proves that the two models established based on ANFIS can track the MR damper force for different amplitude and frequency. In addition, we can predict the voltage required to be applied to produce a known force determined from control law with high performance.

Keywords: ANFIS, Forward fuzzy model, Inverse fuzzy model, Magnetorheological damper.

1. Introduction

Magnetorheological damper is a semi-active device that requires minimal power (20-50 watt); this device can operate with battery eliminating the need for a large power source. Because the device forces are adjusted by varying the strength of the magnetic field, no mechanical valves are required making a highly reliable device [1]. MR dampers are used in various applications, which have been and continue to be developed. These dampers mainly used in industry such as heavy motor damping, vehicle suspension [2]-[7], seismic protection of building during earthquakes [1], [8]-

[18], and the operator seat damping in the construction vehicle [19]. Other fields use MR damper is the military and commercial helicopter cockpit seat. Another application humanitarian is the prosthetic leg. Its decrease the shock delivered to the patient's leg when he is jumping [20], [21].

Parametric models such as Bingham's model, extended Bingham, and Bouc-Wen, which numerically tractable and has been used extensively for modeling hysteretic systems [22]. Regardless of the advantage of

this model, the force-velocity relation within the low velocity range does not closely reproduce the measurement results obtained from a typical MR damper [23]. So, based on extensive experiments, Spencer J. et al. [24], [25] proposed a modified Bouc-Wen model, which avoided most of the problems observed in the other models. This model could keep up all advantages exhibited by the Bouc-Wen model. The parameters associated with this model were found through matching the simulated data with experimental data obtained in a variety test.

Several techniques are used for emulating non parametric models such as neural networks, fuzzy logic, neuro-fuzzy and polynomial models. Wang and Liao [26] proposed a forward MR damper model, which consists of a recurrent neural-network in which the output is delayed and fed back to the input layer. The Levenberg- Marquardt algorithm is the selected to be training method and the training data was obtained from the phenomenological model developed by Spencer J. et al. [24], [27]. The results show that the predicted damping



force approximates the target data except at low-frequency. However, as mentioned by the authors, these are still a preliminary work and further research is required before practical applications. [28]- [30] presented a forward neuro fuzzy model for large-scale and 30 ton MR damper. The reminder of this paper is organized as follows: Section (2) focuses on the model analysis of MR damper. Sections (3) describes and discusses the forward fuzzy model. While, section (4) describes and discusses inverse fuzzy model. Finally, the conclusion is presented in section (5).

2. Model analysis of MR damper

MR damper is consisting of the main hydraulic cylinder which, houses the piston, the magnetic circuit, and accumulator. The cylinder is filled with MR fluid which, composed of micron-sized magnetically polarizable particles dispersed in a carrier medium such as water, mineral or synthetic oil. Typically, it contains 20 to 40% by volume of relatively pure carbonyl iron with 3 to 5 microns in diameter [31]. MR fluid is normally a free flowing viscous fluid, but the presence of a magnetic field causes the particles to form chains and increase the fluid viscosity, until it becomes a semi-solid (Figure 1.a).

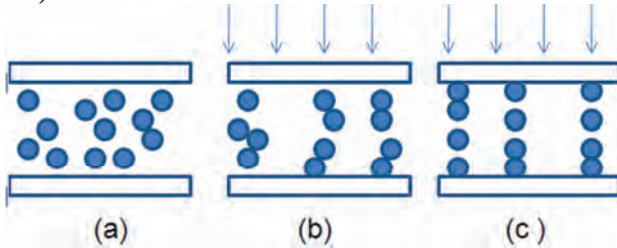


Figure 1.a: Operation of MR fluid (a) no magnetic field (b) and (c) with increasing magnetic field.

The modified Bouc-Wen model is shown in Figure.1.b. It is consisted of mechanical elements such as springs and dashpots to emulate the device behavior.

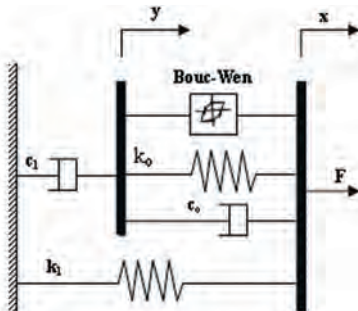


Figure 1.b: Modified Bouc-Wen Model.

The model is described by the following seven nonlinear differential equations [25]:

$$F = c_1 \dot{y} + k_1(x - x_o) \quad (1)$$

where

F : The damping force.

c_1 : The viscous damping for force roll-off at low velocities.

k_1 : The accumulator stiffness.

x : The relative displacement.

x_o : The initial deflection of the accumulator gas spring.

The internal displacement of the damper y and the evolutionary variable z are governed by the following equations:

$$\dot{y} = \frac{1}{c_o + c_1} \{ \alpha z + c_o \dot{x} + k_o(x - y) \} \quad (2)$$

Reference [32] described the restoring forces with hysteresis in the following form:

$$\dot{z} = -\gamma |\dot{x} - \dot{y}| z^{n-1} |z| - \beta (\dot{x} - \dot{y}) z^n + A(\dot{x} - \dot{y}) \quad n \text{ even (3-a)}$$

$$\dot{z} = -\gamma |\dot{x} - \dot{y}| z^n - \beta (\dot{x} - \dot{y}) |z|^n + A(\dot{x} - \dot{y}) \quad n \text{ odd (3-b)}$$

where

c_o : The viscous damping at large velocities.

k_o : The viscous damping for force roll-off at low velocities.

$\dot{x}$: The piston velocity.

α : The Bouc-Wen parameter describing the MR fluid yield stress.

n : Constant.

The adjustment of hysteresis parameters α , β , and A determines the linearity in the unloading region as well as the transition smoothness from the pre-yield to the post-yield region. The parameters γ , β , A , n and k_1 are considered fixed and the parameters c_o , c_1 and α are assumed to be a function of the applied voltage u .

$$\alpha = \alpha_a + \alpha_b u \quad (4)$$

$$c_1 = c_{1a} + c_{1b} u \quad (5)$$

$$c_o = c_{oa} + c_{ob} u \quad (6)$$

If u_c is the command voltage sent to the current driver, the dynamics involved in the MR fluid reaching rheological equilibrium is modeled by the first order filter.

$$\dot{u} = -\eta(u - u_c) \quad (7)$$

where η is the time constant.

Dyke [17] shows the response of the model is dominated by the input velocity and is extremely sensitive to change in velocity. This property causes the model behave erratically when there is a sudden



change in the velocity. Then a Low Pass Filter (LPF) is added to improve the ability of the model to deal with noisy displacement data.

The dynamics of the MR damper with LPF becomes:

$$\dot{y} = \frac{1}{c_o + c_1} \{ \alpha z + c_o v + k_o (x - y) \} \quad (8)$$

$$\dot{z} = -\gamma |v - y| z^{n-1} |z| - \beta (v - y) z^n + (v - y) \quad (9)$$

$$\dot{w} = A_f w + B_f \dot{x} \quad (10)$$

$$v = C_f w$$

$$A_f = \begin{bmatrix} 0 & 1 \\ -\omega_c^2 & \sqrt{2\omega_c} \end{bmatrix} \quad B_f = \begin{bmatrix} 0 \\ 1 \end{bmatrix} \quad C_f = \begin{bmatrix} \omega_c^2 & 0 \end{bmatrix}$$

Where A_f , B_f and C_f are system matrices of LPF, w is the state of the filter and v is the filtered velocity, ω_c is the cut off frequency. The 14 parameters which characteristic the MR damper are given in Table I.

Table I: Parameters for MR-Damper

Parameter	Value	Parameter	Value
c_{0a}	21.0 N.sec/cm	α_a	140 N/cm
c_{0b}	3.50 N.sec/cm.V	α_b	695 .sec/cm.V
k_o	46.9 N/cm	β	363 cm ⁻²
c_{1a}	283 N.sec/cm	β	363 cm ⁻²
c_{1b}	2.95 N.sec/cm.V	A	301
k_1	5.00 N/cm	n	2
x_0	14.3 cm	η	190 sec ⁻¹

3. Forward fuzzy model

Neuro Fuzzy model is a non parametric model for emulating MR damper's behavior. It is faster than the mathematical model for keeping the error relatively small. The neuro fuzzy is an application of ANFIS. ANFIS model using neuro adaptive learning techniques, which are similar to those of neural networks, were originally presented by Jang [33], [34].

3.1 Anfis Architecture

The architecture of a two-input two-rule ANFIS, as example is shown in Figure 2.

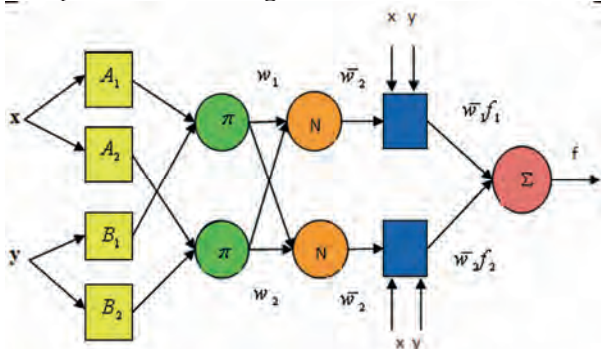


Figure 2: ANFIS architecture.

The ANFIS has five layers [33], in which node functions of the same layer have the same function type as described below: (Note that o_{ij} denotes the output of the i th node in the j th layer).

■ Layer 1: Every node i in this layer is an adaptive node with a node output defined by

$$O_{1,i} = \mu_{A_i}(x), i = 1, 2 \text{ or } O_{1,i} = \mu_{B_{i-2}}(y), i = 3, 4 \quad (11)$$

where x (or y) is the input to the node and A_i (or B_{i-2}) is the fuzzy set associated with this node

$$\mu_{A_i}(x) = \frac{1}{1 + \left[\left(\frac{x - c_i}{a_i} \right)^2 \right]^{b_i}} \quad (12)$$

Where, $\{a_i, b_i, c_i\}$ is the parameter set premise parameters.

■ Layer 2: Every node in this layer is a fixed node labeled Π , which multiplies the incoming signals and outputs the T-norm operator result, e.g.

$$O_{2,i} = w_i = \mu_{A_i}(x) \times \mu_{B_i}(y), i = 1, 2. \quad (13)$$

Each node output represents the firing strength of a rule.

■ Layer 3: Every node in this layer is a fixed node labeled N

$$O_{3,i} = \bar{w}_i = \frac{w_i}{w_1 + w_2}, i = 1, 2. \quad (14)$$

Outputs are called normalized firing strengths.

■ Layer 4: Every node i in this layer is an adaptive node with a node function

$$O_{4,i} = \bar{w}_i f_i = \bar{w}_i (p_i x + q_i y + r_i) \quad (15)$$

where $\bar{w}_i$ is the output of layer 3 and $\{p_i, q_i, r_i\}$ is the parameter set consequent parameters.

■ Layer 5: The single node in this layer is labeled Σ , which computes the overall output as the summation of incoming signals.

$$O_{5,1} = \sum_i \bar{w}_i f_i = \frac{\sum_i w_i f_i}{\sum_i w_i} \quad (16)$$



3.2 Training Forward model

Given input/output data sets, ANFIS constructs Fuzzy Inference System (FIS) whose membership function parameters are adjusted using a back propagation algorithm. The size of input-output data must be large enough and cover all frequency and amplitude ranges to fine tuning the membership function. Three inputs corresponding parameters namely; displacement, velocity, voltage and the MR damper force was the output, which used for the ANFIS training and checking data to build the forward fuzzy model. The training and checking data are obtained from simulating the mathematical Equations (1), (4)-(10) into MATLAB program. The displacement inputs generated using chirp signals with amplitude 4 cm and frequency between 0–3 Hz. The voltage inputs generated using Gaussian white noise ranges from 0–3 V with frequency 0–3 Hz. Our simulation was based on off-line learning only. The data were collected for 20 second and sampled at 1000 Hz, which generate 20000 points of data.

The odd sequence numbers were chosen to be the training data (10000 points) while the even sequence numbers were used as checking data (other 10000 points). The power spectrum density for input and output data are shown in Figure 3.

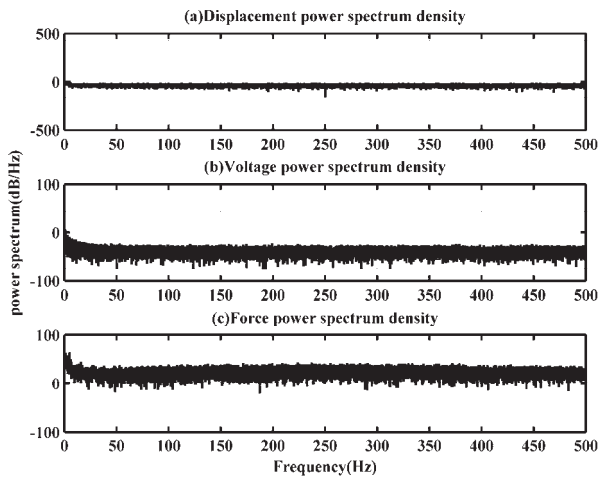


Figure 3: Power spectrum density of displacement, voltage and force

3.3 Validation of forward model

To validate the accuracy of the forward model, five sets of data are generated from mathematical equations to determine the target forces. The first and second sets are sinusoidal excitation with 3cm amplitude, two different frequency 2.5 Hz for data set I and 1 Hz for data set II and held for three constant voltage 0V, 1V, and 2.25 V. Figures (4) and (5) show the comparison between time history of target and predicted forces and force-velocity relationship. It can be seen that the predicted force can track the target force very well for both 0V and 2.25 V

but in the case of constant 1 V the hysteresis behavior predicted not coincide with the target.

The other three validation data sets are chirping displacement with step, pulse, and sinusoidal voltage. For this validation sets the voltage time history is plotted in Figure (6). Figures (7)-(9) show the comparisons between the target and predicted forces versus time and velocity. The predicted forces and hysteresis behavior are matched excellently with the target for a varying voltage with time (the case of semi-active). The accuracy of the neuro fuzzy forward model is proven theoretically by using the error analysis which described by Spencer et al. [25]. The error analysis for all data sets are shown in Table II. E_t , E_v , and E_a are the normalized model error as a function of time, displacement and velocity. From this table, the results show that the force predicted by forward fuzzy model can predict the target force for all cases very well.

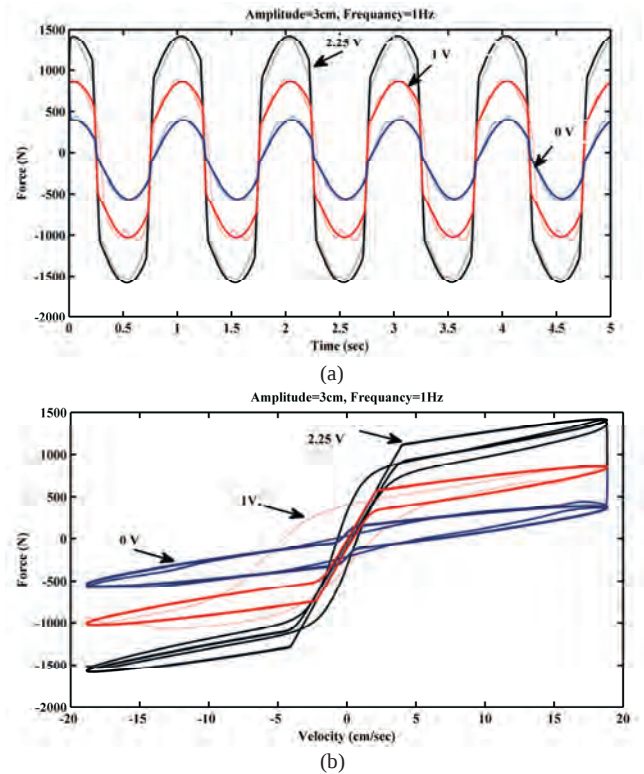


Figure 4: Comparison between fuzzy and mathematical model for data set I: (a) force-time history ; (b) force-velocity relationship.

Table II: Error norm for validation of the forward fuzzy model

Data Sets	Disp.	Volt	E_t	E_v	E_a
I	$3\sin(5\pi t)$	0	0.0004	0.0027	0.0007
I	$3\sin(5\pi t)$	2.25	0.0002	0.0017	0.0004
II	$3\sin(2\pi t)$	1	0.00003	0.0003	0.0003
II	$3\sin(2\pi t)$	2.25	0.0003	0.0012	0.0002
III	Chirp(0-3)Hz	step	0.0002	0.0021	0.0005
IV	Chirp(0-3)Hz	pulse	0.0077	0.0673	0.008
VI	Chirp(0-3)Hz	$1+0.5\sin\pi t$	0.0007	0.0004	0.0008



Table III: comparison between the error norm for the present and previous work [29]

Disp.	Volt	E_t		E_v		E_a	
$2\sin 4\pi t$	0	0.078	0.037	0.271	0.136	1.244	0.533
$2\sin 4\pi t$	2.25	0.027	0.040	0.098	0.086	0.541	0.875
$0.5\sin \pi t$	0	0.013	0.471	0.016	0.441	0.001	0.904
$0.5\sin \pi t$	2.25	0.011	0.235	0.014	0.264	0.001	0.365
GWN	GWN	0.001	0.049	0.003	0.134	0.001	0.674
0-2Hz	0-2Hz						

● The present work ● The previous work[29]

A comparison between the present ANFIS forward model and forward model created by other [29] are shown in Table III. It is show that the values of calculated error norm are very small compared with the same set of data and the same time span.

4. Inverse fuzzy model

Control of the MR damper is very important for all applications. Because the MR damper force cannot be directly controlled, then we can control the applied voltage which, can be directly controlled to operate the MR damper. To achieve that, we must build an inverse model. It predicts the voltage required to be applied to the current driver to produce a desired force with known displacement and velocity.

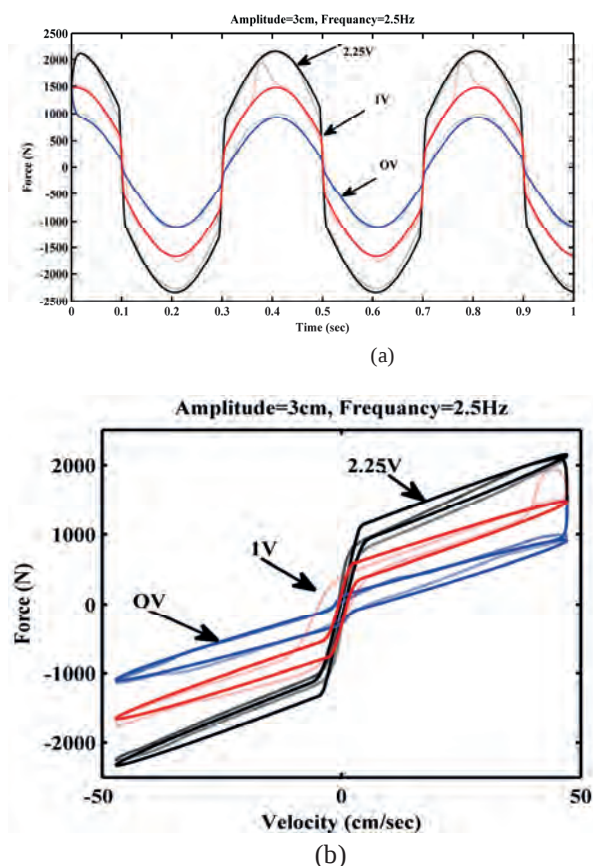


Figure 5: Comparison between fuzzy and mathematical model for data set II: (a) force-time history, (b) force-velocity.

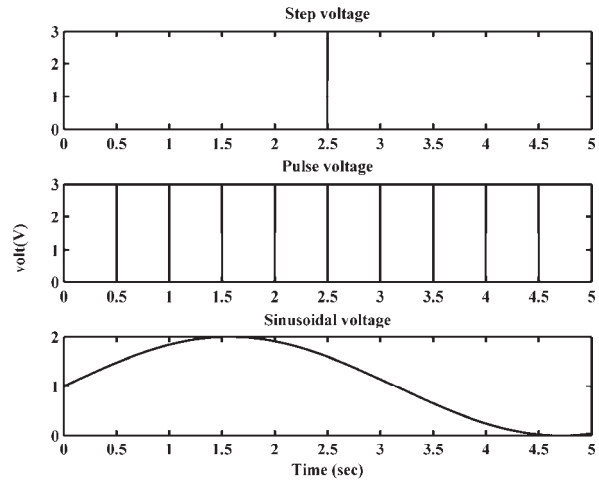


Figure 6: Voltage time history for data sets III, IV, and VI.

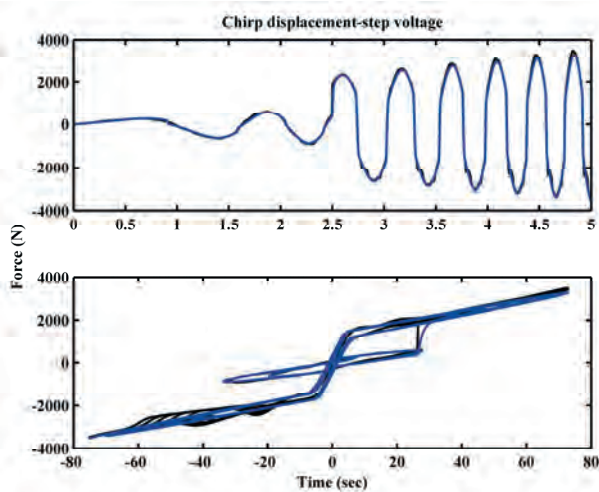


Figure 7: Comparison between fuzzy (black) and mathematical models (blue) for data set III.

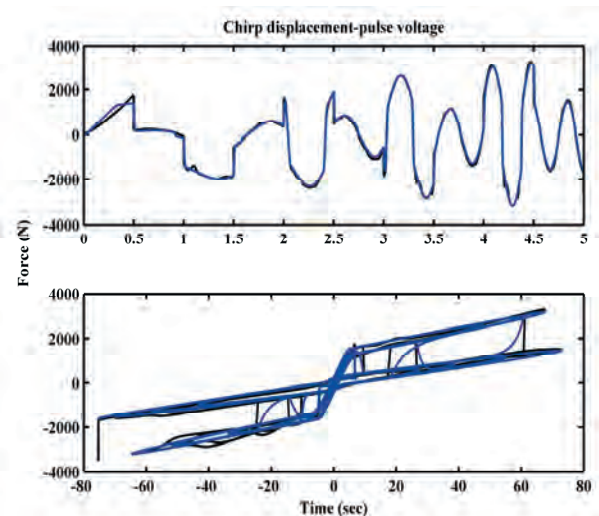


Figure 8: Comparison between fuzzy (black) and mathematical models (blue) for data set IV.



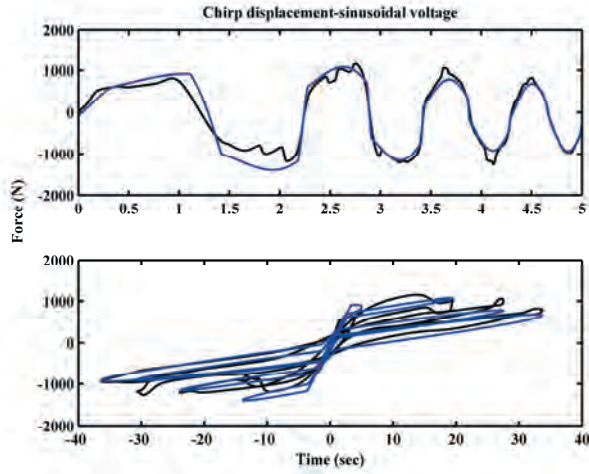


Figure 9: Comparison between fuzzy (black) and mathematical models (blue) for data set VI.

Table IV Input selection of five fuzzy models

Type	$f(k)$	$v(k-1)$	$x(k)$	$\dot{x}(k)$	$f(k-1)$
Type 1					
Type 2					
Type 3					
Type 4					
Type 5					

A quick way to do input selection for inverse model is proposed by Roger Jang [35] based on the assumption that ANFIS model with the smallest root mean square error (RMSE) after one epoch of training has a great potential of achieving a lower RMSE when give more epochs of training. In our study, we tried five models with a different combination of four inputs and train them with a single pass of the least square method (Table IV). Figure 10 shows that the smallest training error is achieved when the inputs are force, displacement, velocity at the current time step, and voltage from previous time step to produce output voltage at the current time step. Then the selected inputs are trained using the hybrid learning rule to tune the membership functions as well (see Figure 11-a). The inverse model is used to predict the requirement voltage, which input to the MR damper model to produce the desired force as illustrated in application phase (Figure 11-b).

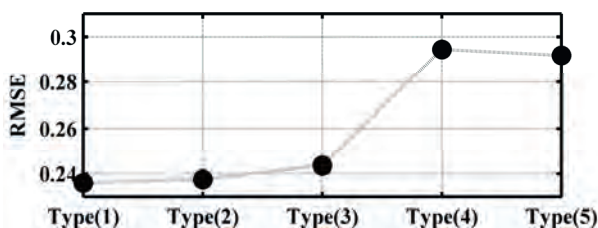


Figure 10: Input selection for inverse fuzzy MR model.

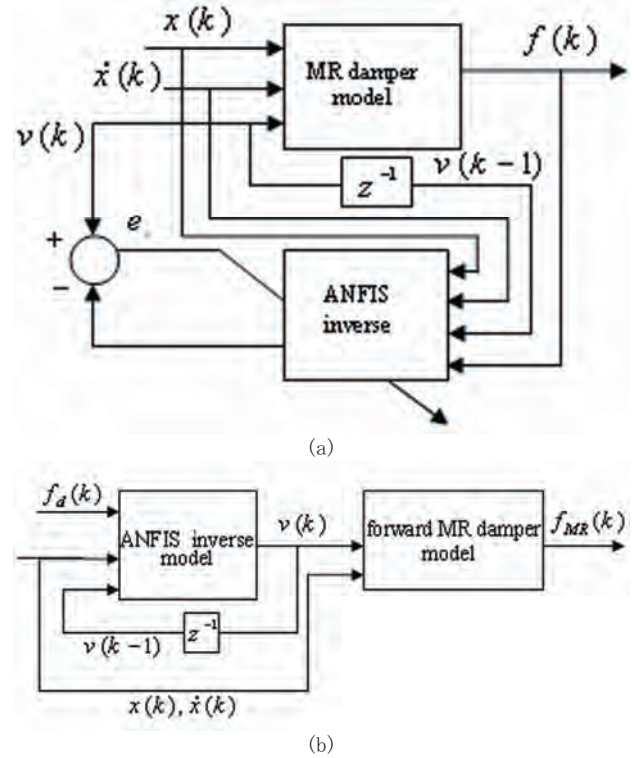


Figure 11: Block diagram for inverse MR damper dynamics: (a) Training phase; (b) application phase.

In order to investigate the inverse dynamics fuzzy model, a variety of investigation data are applied. The investigation data has been collected for five seconds and sampling at 2000 Hz. The input is the displacement of 2.5 Hz sinusoidal excitation with an amplitude 2 cm.

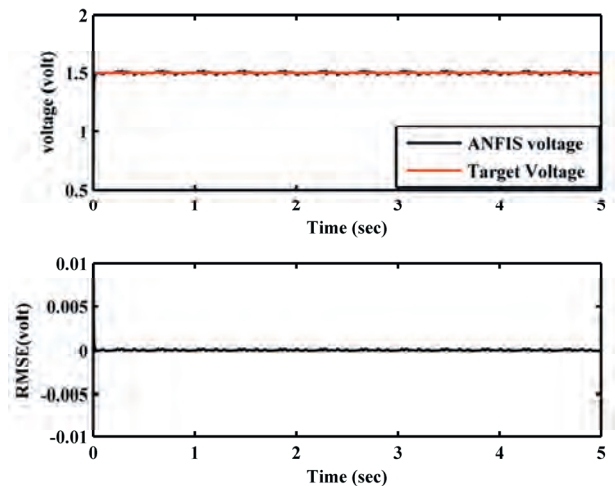


Figure 12: Comparison between predicted and target voltage and the root mean square errors for constant 1.5V.

The target voltage is varying with time from 0.5 to 1.5 V for three tests and constant with 1.5 V for fourth test. Figures (12)-(15) compare the predicted voltage produced by the inverse fuzzy model with the target voltage and present the root mean



square error (RMSE) between them. Figure (12) and (13) compare the predicted and target when the voltage is a constant value of 1.5 V and sinusoidal ($1+0.5 \sin \pi t$). It is seen that the RMSE is about zero and the predicted voltage is match the target with high-efficiency.

Figure (14) and (15) compare the predicted and target when the voltage is step and sawtooth with an amplitude of 1.5 V. It is shown that the predicted voltage is coincided with the target except at the peaks with RMSE 0.035 and 0.02 respectively.

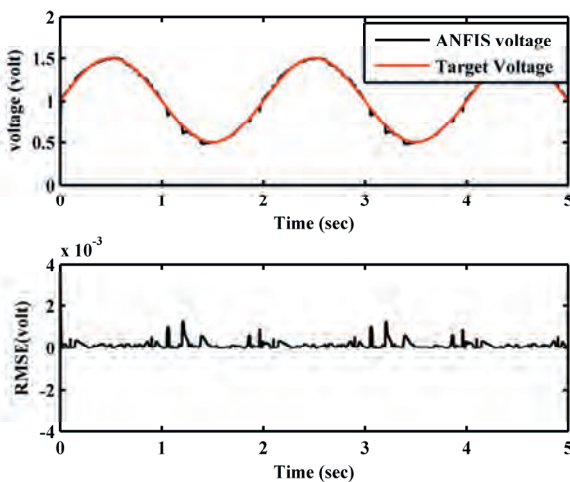


Figure 13: Comparison between predicted and target voltage and the root mean square errors for ($1+0.5 \sin \pi t$) V.

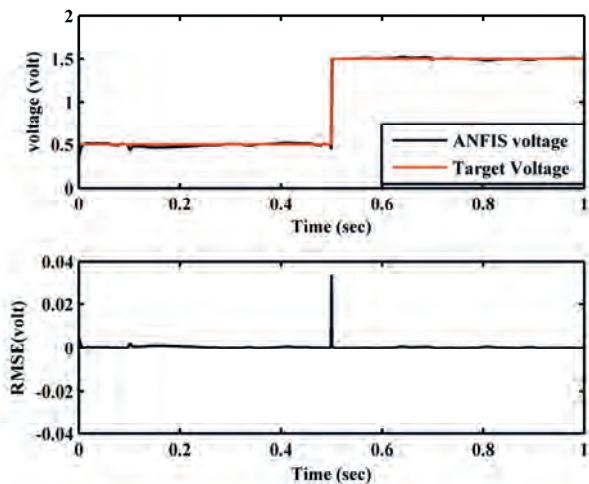


Figure 14: Comparison between predicted and target voltage and the root mean square errors for step voltage.

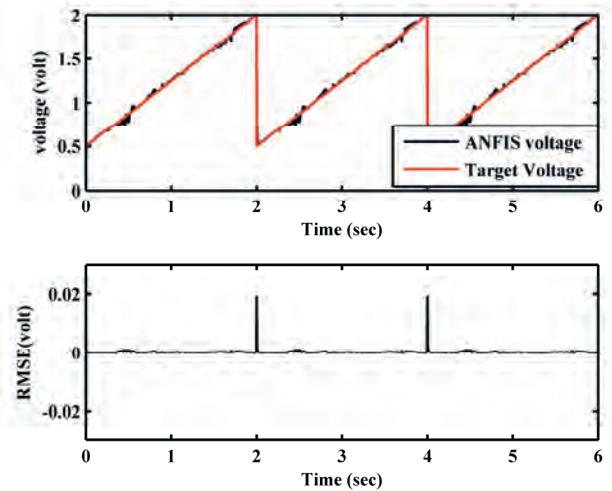


Figure 15: Comparison between predicted and target voltage and the root mean square errors for sawtooth with amplitude 2 V.

5. Conclusion

The magnetorheological damper is a versatile device, which can be used in many applications. The identification of MR damper is definitely difficult due to its nonlinear dynamics. In this paper, ANFIS technique is applied to build the forward and inverse fuzzy MR damper models. The dynamic behavior of the damper is described by the membership function. Data used for training the model is generated from numerical simulation of nonlinear differential equations. The same data are used to train the inverse model, which predicts the voltage required to produce the desired force. Validation results show that the proposed forward fuzzy model can predict the damping force accurately. Furthermore, the inverse dynamic model can generate the target voltage when the MR damper is used in any control system.

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Biography



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CONTROL OF CASCADED MULTILEVEL INVERTER BY USING CARRIER BASED PWM TECHNIQUE AND IMPLEMENTED TO INDUCTION MOTOR DRIVE

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Abstract

Multilevel power conversion technology is a very rapidly growing area of power electronics with good potential for further development. Today, it is hard to connect a single power semiconductor switch directly to medium voltage grids. For these reasons, a new family of multilevel inverters has emerged as the solution for working with higher voltage levels.

A Carrier Based Pulse Width Modulation (CBPWM) Scheme for multilevel inverters is proposed. The proposed PWM scheme generates the inverter leg switching times from the sampled reference phase voltage amplitudes and centers the switching times for the middle vectors, in a sampling interval, as in the case of conventional space vector PWM (SVPWM). The SVPWM scheme, presented for multilevel inverters, can also work in the over modulation range, using only the sampled amplitudes of reference phase voltages. The sinusoidal PWM technique for various modulation indices, Carrier Based PWM technique for the cascaded three-level and five-level inverter fed induction motor are analyzed and discussed. The performance of induction motor is studied in various aspects for various modulating techniques and the results are compared and tabulated. An induction motor model is developed using Simulink and it is fed from a three phase three phase five-level cascaded inverter. The above said sampled amplitudes of reference phase voltages based SVPWM technique is implemented using MATLAB/SIMULINK.

Key words Cascaded Multilevel Inverter, PWM Techniques, Carrier Based PWM, Induction Motor, and Total Harmonic Distortion.

1. Introduction

The general structure of the multilevel converter is to synthesize a near sinusoidal voltage from several levels of dc levels of dc voltages, typically obtained from capacitor voltage sources. As the number of levels increases, the synthesized output waveform has more steps, which produce a staircase wave that approaches a desired waveform [4][5]. Also, as more steps are added to the waveform, the harmonic distortion of the output wave decreases, approaching zero as a number of levels increases. As the number of levels increases, the voltage that can be spanned by summing multiple voltage levels also increase. The output voltage during the positive half-cycles can be found from

$$v_{a0} = \sum_{n=1}^m E_n SF_n$$

Where SF_n is the switching or control function of n th node it takes a value of 0 or 1. Generally, the capacitor terminal voltages $E_1, E_2, \dots$ all have the same value E_m . Thus, the peak output voltage is $v_{a0}(\text{peak}) = (m-1)E_m = V_{dc}$. To generate an output voltage with both positive and negative values, the circuit topology [3] has another switch to produce the negative part v_{ob} so that.

The multilevel inverters can be classified into three types [1, 2].

1. Diode-clamped multilevel inverter.
2. Flying-capacitors multilevel inverter.
3. Cascade multilevel inverter.

Among the above specified three topologies Cascade Multilevel inverter has the features as follows:



For real power conversions from ac to dc and then dc to ac, the cascaded inverters needs separate dc sources. The structure of separate dc sources is well suited for various renewable energy sources such as fuel cell, photovoltaic, and biomass. Connecting dc sources between two converters in a back-to-back fashion is not possible because a short circuit can be introduced when two back-to-back converters are not switching synchronously [6] [7].

The advantages of a cascade multilevel H-bridge inverter with s separate dc sources per phase is as follows [1, 2]

- The series structure allows a scalable, modularized circuit layout and packaging since each bridge has the same structure
- Requires the least number of components considering there are no extra clamping diodes or voltage balancing capacitors[8]
- Switching redundancy for inner voltage levels is possible because the phase voltage output is the sum of each bridge's.

Table 1 compares the component requirements per phase leg among the three multilevel converters. All devices are assumed to have the same voltage rating, but necessarily the same current rating.

Converter type	Diode-clamp	Flying-capacitors	Cascaded inverters
Main switching devices	$(m-1) \times 2$	$(m-1) \times 2$	$(m-1) \times 2$
Main diodes	$(m-1) \times 2$	$(m-1) \times 2$	$(m-1) \times 2$
Clamping diodes	$(m-1) \times (m-2)$	0	0
Dc bus capacitors	$(m-1)$	$(m-1)$	$(m-1)/2$
Balancing capacitors	0	$(m-1) \times (m-2) / 2$	0

Table 1 Comparisons of component requirement per leg of three multilevel inverters [2]

The two most widely used PWM schemes for multilevel inverters are the Modified Reference with triangular carrier PWM technique [11] and Modified Carrier with Sine reference PWM technique. In case of first the reference waves can be apart from the sine wave i.e. like trapezoidal, step, third harmonic injected or space vector PWM etc., and in case of next the carrier wave can be apart from the regular triangular wave i.e triangle with different phase dispositions and u-type carriers with phase dispositions.



In this paper, the second technique i.e Modified Carrier with Sine reference PWM technique and compared with the new proposed modulation technique called as Modified Reference with Modified Carrier modulation technique. In the proposed case offset voltage injected reference is compared with the u-type carrier signals with different phase dispositions. This obtained gate pulses are applied to a Cascaded Five level inverter when it feeds to an induction motor. This proposed modulation technique called as Modified SVPWM, because the reference wave in the case of SVPWM looks similar with the proposed one, but in performance wise, implementation wise the proposed one has no sector identification with vector based technique and cumbersome calculations. This paper has carried out the simulation results for Sinusoidal PWM with different carrier waves and Modified Reference with Modified Carrier PWM techniques when implemented to a Cascaded five level inverter and feed a induction motor. The comparative results are taken at output voltage of inverter and performance tested in terms of THD.

2. Modulation techniques

Sinusoidal PWM

Pulse width modulation control is most widely used method of controlling the modulation depth of inverters, including the multilevel family. A significant amount of research has been published on the various ways of Implementing PWM control. The focus here is on carrier-based sinusoidal PWM schemes for controlling a cascaded multilevel inverter.

The SPWM schemes are more flexible and simpler to implement, but the maximum peak of the fundamental component in the output voltage is limited to 50% of the DC link voltage [10], and the extension of the SPWM schemes into the over-modulation range is difficult. The SPWM technique, when applied to multilevel inverters, uses a number of level-shifted carrier waves to compare with reference phase voltage signals.

SPWM is the most widely used PWM technique for inverters of two and more levels. The basic principle of the SPWM used in two level inverters is explained here. The reference signal (V_r) which generally is sinusoidal is compared with the high frequency triangle wave (V_c) of constant amplitude. When $V_r > V_c$, the PWM output will be high (state +1), and output will be low (state -1) for $V_r < V_c$.

In general, the modulation index is defined as the ratio of the magnitude of the reference signal to that of the magnitude of the carrier signal. The modulation index is $m = V_r / V_c$, Where V_r is the magnitude of the sinusoidal reference signal waveform and V_c is the magnitude of the triangular carrier signal. By varying



V_r and keeping V_c constant, the modulation index can be changed. By changing the modulation index, the amplitude of the fundamental component of the output can be varied. For three phase inverters, the same carrier signal V_c is used for all the three phases and three reference signals which are phase displaced by $2\pi/3$ radians are used for each of the phases.

The parameters of the modulation process are : The ratio $P = \omega_c/\omega_m$, The ratio $M = A_m$ and the angle ϕ of displacement existing between the sinusoidal reference and the first positive triangular carrier signals

We also have other degrees of freedom. In fact, the phase displacement between two contiguous triangular carriers is free. We consider only three very simple dispositions that seem the most interesting:

- All the carriers are phase disposition (PD)
- All the carriers above the zero value reference are in phase among them but in opposition with those below (POD)
- All the carriers are alternatively in opposition Disposition (APOD)

In this paper both triangular and u-type carriers are analyzed. The cascaded five-level inverter with R-load is simulated and the results are explained below:

Phase Disposition (PD)

In this method carriers are the same in frequency, amplitude and phases, but they are just different in DC offset to occupy contiguous bands as shown in Figure 1. The carriers are in phase across all the bands. For this technique, significant harmonic energy is concentrated at the carrier frequency, but since it is co-phase component, it does not appear in the line-to-line voltage.

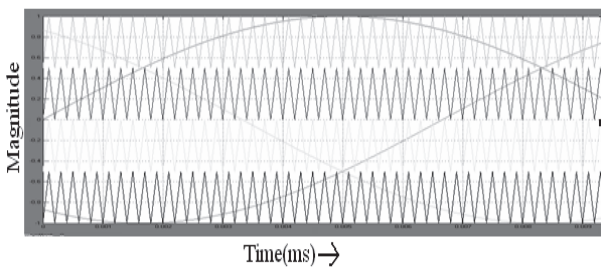


Figure. 1: Generation of gate pulses with triangular carriers

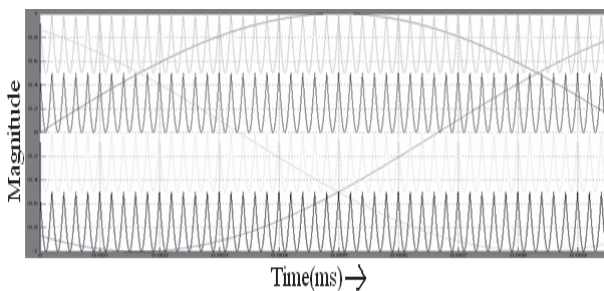


Figure. 2: Generation of gate pulses with U-type carriers

Figure. 1 shows the generation of gate pulses with triangular carriers of cascaded five-level inverter with R-load. In this four triangular carriers are selected on the basis of the formula $m-1$, i.e $5-1=4$. Figure. 2 shows the generation of gate pulses with u-type carrier

Phase Opposition Disposition (POD)

Carriers in this method are the same in frequency and amplitude but they are again different in phase and DC offset as the case of APOD. But in this method carriers above the reference zero point are out of phase with those below that by 180° as shown in Figure.3

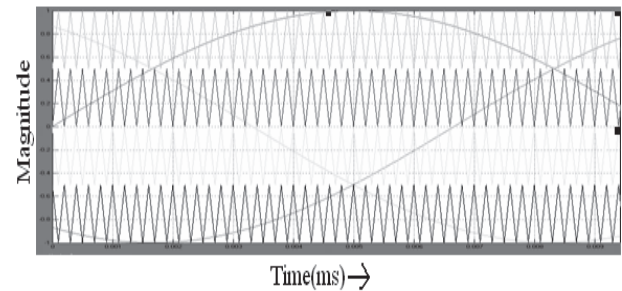


Figure. 3: Generation of gate pulses with triangular carriers

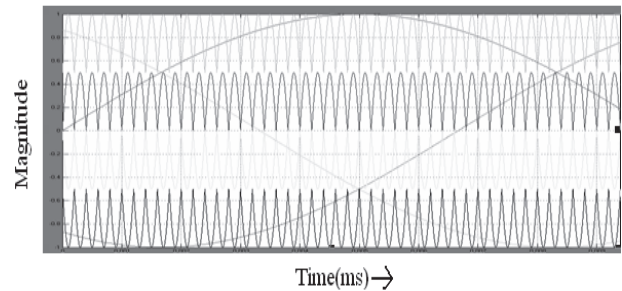


Figure. 4: Generation of gate pulses with U-type carriers

Alternative Phase Opposition Disposition (APOD)

In this method carriers have the same frequency and the same amplitude but they are different in their DC offset and phases as shown in Figure.5. In this method carriers are phase shifted by 180° , so this method uses two degrees of freedom of carriers namely their DC-offset and phases.

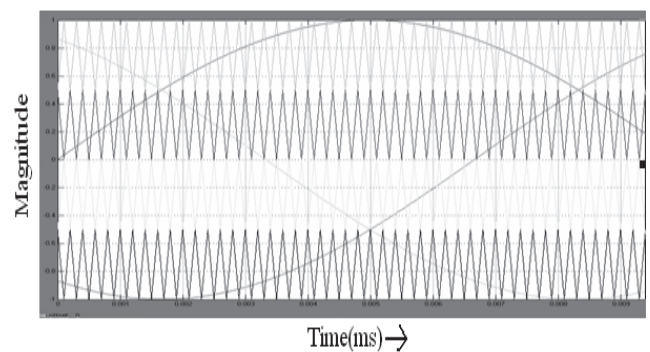


Figure. 5: Generation of gate pulses with triangular carriers



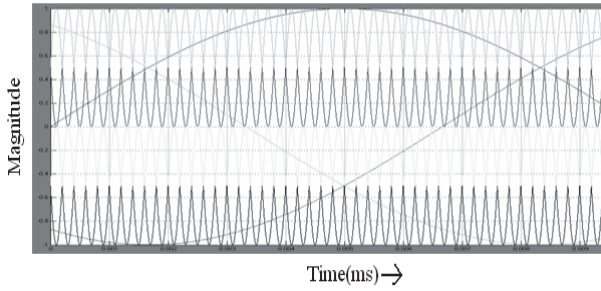


Figure. 6: Generation of gate pulses with U-type carriers

3. Modified SVPWM

In SVPWM scheme gives a more fundamental voltage and better harmonic performance compared to the SPWM schemes [12]. The maximum peak of the fundamental component in the output voltage obtained with space vector modulation is 15% greater than with the sine-triangle modulation scheme. But the conventional SVPWM scheme requires sector identification and look-up tables to determine the timings for various switching vectors of the inverter, in all the sectors. This makes the implementation of the SVPWM scheme quite complicated the over modulation range, extending up to six-step operation. It has been shown that, for two-level inverters, a SVPWM-like performance can be obtained with a SPWM scheme by adding a common mode voltage of suitable magnitude, to the sinusoidal reference phase voltage. A simplified method, to determine the correct offset times for centering the time durations of the middle inverter vectors, in a sampling interval, is presented, for the two-level inverter. The inverter leg switching times are calculated directly from the sampled amplitudes of the reference three-phase voltages with considerable reduction in the computation time [13].

To obtain the maximum possible peak amplitude of the fundamental phase voltage, in linear modulation, a common mode voltage, $V_{offset1}$, is added to the reference phase voltages, where the magnitude of $V_{offset1}$ is given by [13]

$$V_{offset1} = \frac{-(V_{max} + V_{min})}{2} \quad (1)$$

In (1), V_{max} is the maximum magnitude of the three sampled reference phase voltages, while V_{min} is the minimum magnitude of the three sampled reference phase voltages, in a sampling interval. The addition of the common mode voltage, $V_{offset1}$, results in the active inverter switching vectors being centered in a sampling interval, making the SPWM technique equivalent to the modified reference PWM technique. Equation (1) is based on the fact that, in a sampling interval, the reference phase which has lowest magnitude (termed the min-phase) crosses the triangular carrier first, and causes the first transition in the inverter switching state. While the reference phase,

which has the maximum magnitude (termed the max-phase), crosses the carrier last and causes the last switching transition in the inverter switching states in a two-level modified reference PWM scheme. Thus the switching periods of the active vectors can be determined from the (max-phase and min-phase) sampled reference phase voltage amplitudes in a two-level inverter scheme. The SPWM technique, for multilevel inverters, involves comparing the reference phase voltage signals with a number of symmetrical level-shifted carrier waves for PWM generation. Because of the level-shifted multi carriers, the first crossing (termed the first-cross) of the reference phase voltage cannot always be the min-phase. Similarly, the last crossing (termed the third-cross) of the reference phase voltage cannot always be the max-phase. The idea behind the proposed scheme is to determine the sampled reference phase, from the three sampled reference phases, which crosses the triangular first (first-cross) and the reference phase which crosses the triangular carrier last (third-cross). Once the first-cross phase and third-cross phase are identified, the principles of offset calculation of equation (1), for the two-level inverter, can easily be adapted for the multilevel modified reference PWM generation scheme. The proposed modified reference PWM technique presents a simple way to determine the time instants at which the three reference phases cross the triangular carriers. These time instants are sorted to find the offset voltage to be added to the reference phase voltages for modified reference PWM generation for multilevel inverters for the entire linear modulation range [14], so that the middle inverter switching vectors are centered (during a sampling interval), as in the case of the conventional two-level modified reference PWM scheme.

For this analysis we have developed two types of carrier based techniques:

- Modified reference with triangular carriers
- Modified reference with modified carriers

As like discussed in SPWM technique the carrier based techniques are [11, 13]:

1. Phase Disposition (PD)
2. Phase Opposition Disposition (POD)
3. Alternate Phase Opposition Disposition (APOD)

Phase Disposition

In this method carriers are the same in frequency, amplitude and phases, but they are just different in DC offset to occupy contiguous bands as shown in Figure 7. The carriers are in phase across all the bands. For this technique, significant harmonic energy is concentrated at the carrier frequency, but since it is co-phase component, it does not appear in the line-to-line voltage.



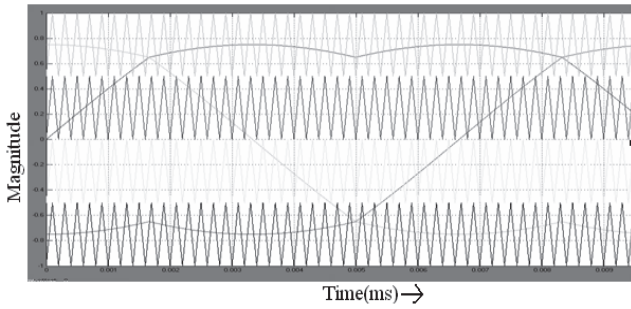


Figure 7: Generation of gate pulses with triangular carriers

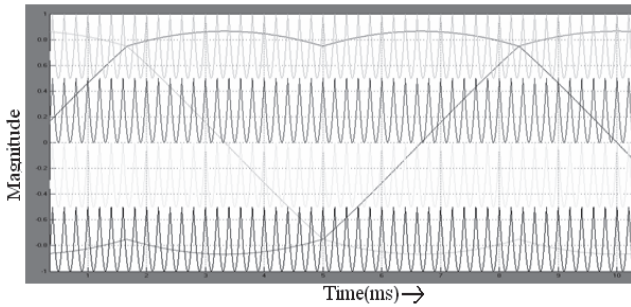


Figure 8 : Generation of gate pulses with U-type carriers

Figure 7 shows the generation of gate pulses with triangular carriers of cascaded five-level inverter with R-load. In this the reference is modified and four triangular carriers are selected on the basis of the formula $m-1$, i.e. $5-1=4$. Figure 3.16 shows the generation of gate pulses with u-type carriers.

Figure 8 is the line-line output voltage of cascaded five-level inverter with R-load and harmonic spectrum. For this the triangular carriers are used. The harmonic spectrum of cascaded five-level inverter with R-load is shown in Figure 9. It shows that the total harmonic distortion is 3.68%.

Figure 10 is the line-line output voltage of cascaded five-level inverter with R-load and harmonic spectrum. For this the u-type carriers are used. The harmonic spectrum of cascaded five-level inverter with R-load is shown in Figure 10. It shows that the total harmonic distortion is 4.68%.

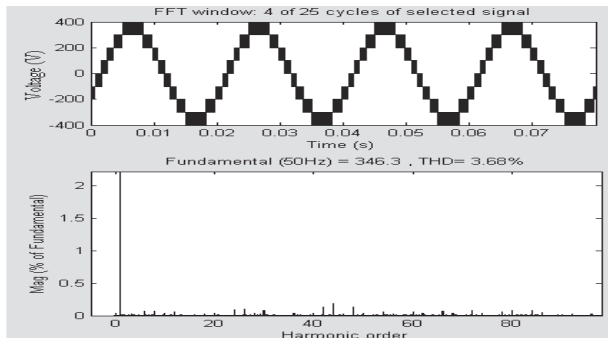


Figure 9: Output line-line voltage and harmonic spectrum with triangular carriers

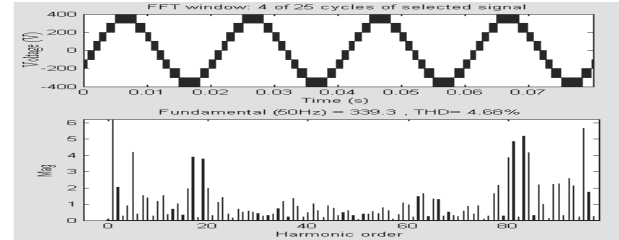


Figure 10: Output line-line voltage and harmonic spectrum with u-type carriers

Phase Opposition Disposition (POD)

Carriers in this method are the same in frequency and amplitude but they are again different in phase and DC offset as the case of APOD. But in this method carriers above the reference zero point are out of phase with those below that by 180° as shown in Figure.11.

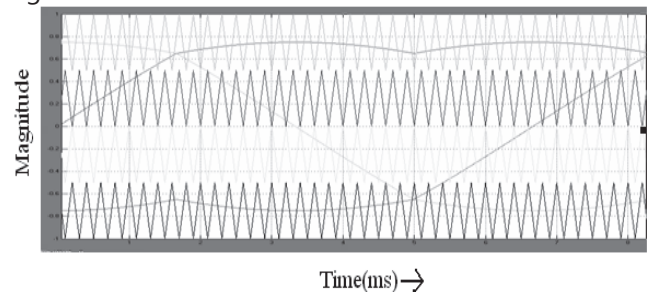


Figure 11: Generation of gate pulses with triangular carriers

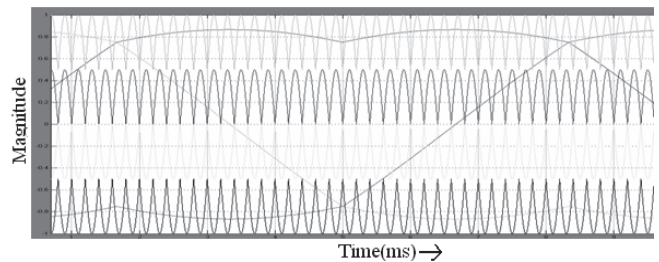


Figure 12: Generation of gate pulses with u-type carriers

Alternative Phase Opposition Disposition (APOD)

In this method carriers have the same frequency and the same amplitude but they are different in their DC offset and phases as shown in Figure.13. In this method carriers are phase shifted by 180° , so this method uses two degrees of freedom of carriers namely their DC-offset and phases.

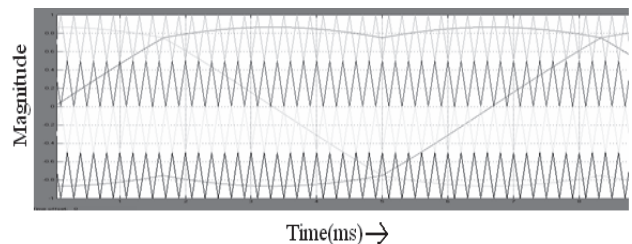


Figure 13: Generation of gate pulses with triangular carriers



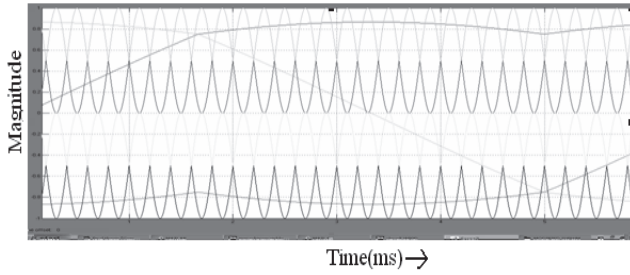


Figure 14: Generation of gate pulses with U-type carriers

Among the discussed techniques, PD technique has less harmonic distortion on line voltages. As it shown PD technique puts the harmonic energy directly into a common mode carrier component so that the harmonics are cancelled in line voltages.

4. Simulation results comparisons

In this section the cascaded multilevel inverter simulation results can be shown with R-load by using the modulation techniques of SPWM and Modified SVPWM. The comparison of simulation results are shown in Table 2 & 3.

Modulation index	Sinusoidal Reference(SPWM)					
	triangular carrier PWM			u-type carrier PWM		
	PD	POD	APOD	UPD	UPOD	UAPOD
M=1	5.08	5.87	12.38	5.97	8.81	12.71
M=0.866	5.26	6.80	12.90	6.30	9.89	12.88
M=0.8	7.36	9.29	17.40	7.13	11.10	14.65

Table 2 The %THD comparison of voltage for SPWM for Cascaded Five-level Inverter

Modulation index	Modified Reference(Modified SVPWM)					
	Triangular Carrier PWM			u-type Carrier PWM		
	PD	POD	APOD	UPD	UPOD	UAPOD
M=1	3.68	5.11	8.57	4.68	4.68	11.41
M=0.866	4.45	5.23	10.38	5.56	6.32	12.56
M=0.8	4.69	6.39	12.25	6.36	7.46	13.01

Table 3 The %THD comparison of voltage for Modified SVPWM for Cascaded Five-level Inverter

5. CC5LI fed induction motor

In this Section the simulation results of cascaded five-level inverter (CC5LI) fed induction motor, line-line voltage of harmonic spectrum, and speed torque characteristics are analyzed and discussed. The modeling of induction motor as follows. The per phase equivalent circuit of the induction motor can be used

for only steady state analysis[15]. The induction motor contains time varying mutual inductances in the transient state which makes the analysis using per phase equivalent circuit and impossible one[16]. The analysis using d-q model can take care of the time varying quantities in the transient state.

The modeling equations in flux linkage form are as follows:

$$\frac{dF_{qs}}{dt} = \omega_b \left\{ v_{qs} - \frac{\omega_e}{\omega_b} F_{ds} + \frac{R_s}{x_{ls}} (F_{mq} + F_{qs}) \right\}$$

$$\frac{dF_{ds}}{dt} = \omega_b \left\{ v_{ds} + \frac{\omega_e}{\omega_b} F_{qs} + \frac{R_s}{x_{ls}} (F_{md} + F_{ds}) \right\}$$

$$\frac{dF_{qr}}{dt} = \omega_b \left\{ v_{qr} - \frac{\omega_e - \omega_r}{\omega_b} F_{dr} + \frac{R_r}{x_{lr}} (F_{mq} - F_{qr}) \right\}$$

$$\frac{dF_{dr}}{dt} = \omega_b \left\{ v_{dr} + \frac{\omega_e - \omega_r}{\omega_b} F_{qr} + \frac{R_r}{x_{lr}} (F_{md} - F_{dr}) \right\}$$

The complete block diagram of cascaded five-level inverter fed induction motor drive is shown in Figure.15. ; in this the cascaded five-level inverter[6] is connected at input supply of the system. Two comparators are used in the firing circuit, one is for 3-phase sine-wave reference signal and the other is for level shifted carriers. By using of this combination total 24 pulses is generated, according this 24 switches are required to generate these pulses.

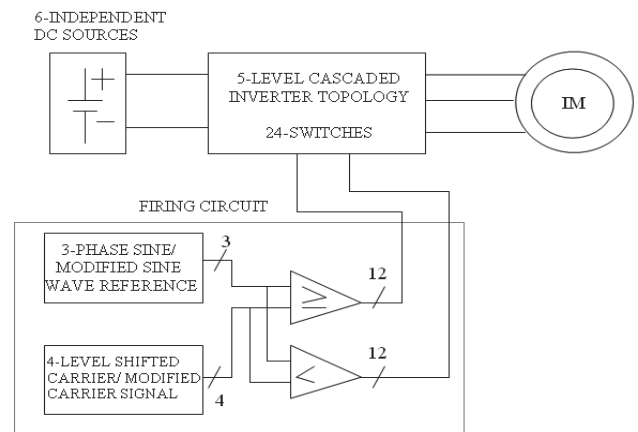


Figure.15: Block diagram of cascaded five level inverter fed induction motor drive

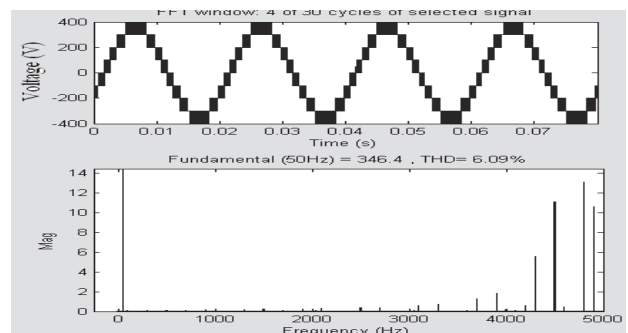


Figure 16: Line-Line output voltage and FFT spectrum for cascaded 5-level inverter fed IM with SPWM



Figure.16 Shows the Line-Line output voltage of cascaded five-level inverter fed induction motor. The harmonic spectrum is also shown with THD 6.09%. The PD modulation scheme of SPWM is used for this simulation.

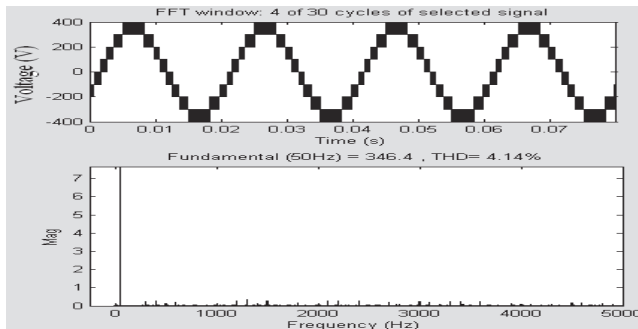


Figure. 17: Line-Line output voltage and FFT spectrum for cascaded 5-level inverter fed IM with modified SVPWM

Figure.17 Shows the Line-Line output voltage of cascaded five-level inverter fed induction motor with no-load. The harmonic spectrum is also shown with THD 4.14%. The PD modulation scheme of Modified SVPWM is used for this simulation

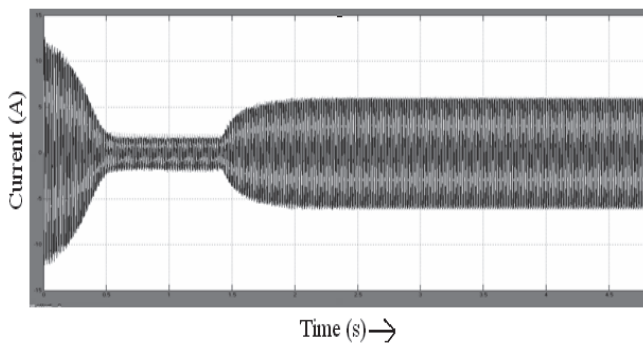


Figure 18 : Stator current of cascaded 5-level inverter fed IM with step load $T_L=7$ N-m

The Output current of induction motor with step-load $T_L=7$ N-m is shown in Figure.18. It shows that the starting current of induction motor is nearly 12A, and after motor reaches the rated speed the current is nearly 1-2Amp, when load is applied the current increases to 5-6 Amp.

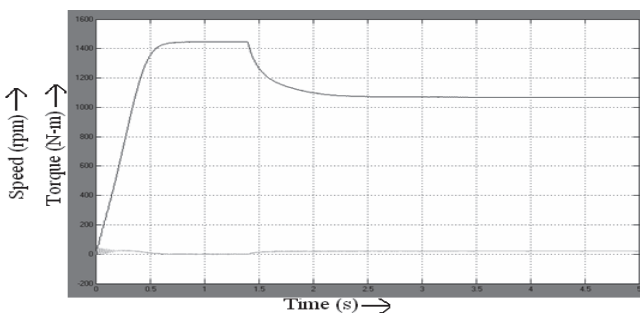


Figure. 19: Speed and torque waveform of cascaded five-level inverter fed with step Load=7 N-m

Figure.19 shows the Speed and torque waveform of an induction motor with step-load of 7 N-m. It shows

that the speed will reaches the rated speed within short time and it retains the rated speed of 1440rpm continuously. When a load of 7-Nm is applied, speed is reduced and retained 1050 rpm. The torque characteristics are also shown in Figure. 19. At the time of starting it shows that the torque oscillates, when the motor reaches its rated speed it comes to constant and when the load is applied on the motor the load torque again increases.

6. Conclusion

The cascaded multilevel inverter has the superior advantages over diode-clamped, capacitor-clamped multilevel inverter topologies. So in this paper, the cascaded multilevel inverter is proposed for implementing of multilevel inverter fed induction motor.

A summary of THD for Cascaded multilevel inverter by using SPWM and Modified SVPWM modulation techniques is presented. Cascaded 5LI are simulated for SPWM, modified reference with triangular carriers and modified reference with modified carrier (U-type). The modeling of induction motor is developed. For this various equations are derived and based on that equations the Simulink block diagram of induction motor is developed.

The simulation results of cascaded five- level inverter fed induction motor with modified SVPWM are analyzed with respect to line-line voltage, the THD spectrum of line-line voltage and torque speed characteristics of induction motor. The %THD of Line-Line output voltage of cascaded five-level inverter fed IM with load 7N-m by using the modified SVPWM Technique is 4.14%.

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Improved Analysis and Simulation of Micro-Technology based on Frequency-Domain Hybrid Reduction Method: Micro-Electromechanical Optical Filter Benchmark

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Abstract

A new method for model reduction is proposed. The reduction technique is based a recently developed methods, namely frequency domain balanced truncation within a frequency bound. This reduction technique is developed for model reduction of very large dynamical systems which are frequently appear in analysis and simulation in Micro-technology as a result of finite element method. The proposed method is of interest for practical model order reduction because in this context it shows to keep the accuracy of the approximation as high as possible without sacrificing the computational efficiency. The method is implemented for simulation and analysis of micro-electromechanical optical filter. Numerical results show the accuracy and efficiency enhancement of the method in comparison to previous techniques in this context.

Keywords: *hybrid method, Dynamical systems, Frequency-Domain Grammian, Model Reduction, Micro-technology.*

1. Introduction

Over the past two decades model reduction has become an ubiquitous tool in a variety of application areas and, accordingly, a research focus for many mathematicians and engineers. The ever-increasing need for accurate mathematical modeling of physical as well as artificial processes for simulation and control leads to models of high complexity. This problem demands efficient computational prototyping tools to replace such

complex models by an approximate simpler models, which are capable of capturing dynamical behavior and preserving essential properties of the complex one, either the complexity appears as high order describing dynamical system or complex nonlinear structure. Due to this fact model reduction methods have become increasingly popular over the last two decades. Such methods are designed to extract a reduced order state space model that adequately describes the behavior of the system in question.

In the field of simulation and analysis in micro and nanotechnology frequently we have to deal with large dynamical systems which are mostly the results of the tools for FEM(finite element method). On the other hand these results are in descriptor structure and usually singular. This problem is motivated the researchers in this context to study model reduction of singular systems. In this paper an accuracy and efficiency enhanced method for model order reduction is presented for model reduction of very large singular system. This method is a two-step framework. In the first step we apply one of the numerically efficient reduction techniques to overcome very large orders problem. Second step in the framework is frequency-domain balanced reduction within frequency bound. This framework provides more efficient and accurate results comparing to the previous counterparts.

The paper is organized as follows. In section 2, we present some system theoretic and mathematical background briefly. Section 3 consists of presenting our method in detail. In section 4 the proposed



methods applied to the model in microtechnology. The conclusion is presented in section 5.

2. Mathematical and System Theoretic Background

In this section some notions and notations in context of model reduction in general and model reduction of singular systems in particular are presented in brief. More details can be found in [1]-[4].

Def. 1. Dynamical singular systems:

$$E\dot{x} = Ax + Bu, \quad y = Cx \quad (1)$$

and

$$\tilde{E}\dot{\tilde{x}} = \tilde{A}\tilde{x} + \tilde{B}u, \quad y = \tilde{C}\tilde{x} \quad (2)$$

are called restricted equivalent systems if there exist non-singular P and Q such that:

$$x = P\tilde{x}, \quad QEP = \tilde{E}, \quad QAP = \tilde{A}, \quad QB = \tilde{B}, \quad CP = \tilde{C}$$

Def. 2. Dynamical singular system (1) is regular if there exist $\alpha \in \mathbb{C}$ such that:

$$\det(\alpha E + A) \neq 0 \quad \text{or} \quad \det(\alpha E - A) \neq 0$$

Following we recall a theorem from [4] which is very important in reduction of dynamical singular systems.

Theorem 1. Every regular singular system has a restricted equivalent dynamical system which is described by the following equations:

$$\dot{x}_1 = A_{11}x_1 + A_{12}x_2 + B_1u \quad (3)$$

$$0 = A_{21}x_1 + A_{22}x_2 + B_2u \quad (4)$$

$$y = C_1x_1 + C_2x_2 \quad (5)$$

In order to get this structure we can apply SVD on E:

$$U^T E V = \text{diag}(\Sigma, 0)$$

and then by coordinate transformation:

$$x = V \begin{bmatrix} x_1 \\ x_2 \end{bmatrix}$$

We have:

$$\begin{bmatrix} A_{11} & A_{12} \\ A_{21} & A_{22} \end{bmatrix} = T U^T A, \quad \begin{bmatrix} B_1 \\ B_2 \end{bmatrix} = T U^T B$$

$$\begin{bmatrix} C_1 & C_2 \end{bmatrix} = C V^T, \quad T = \begin{bmatrix} \Sigma^{-1} & 0 \\ 0 & I \end{bmatrix}$$

This kind of transformation enable us to get non-singular system out of a singular structure (by expressing x_2 in terms of x_1 and eliminating x_2).

3. Frequency-Domain Hybrid Reduction Method

In this section we first present frequency domain balanced truncation method and the related techniques and then the frequency-domain hybrid reduction method for reduction of very large dynamical systems.

3.1 Model order reduction within frequency bound

Consider the following n^{th} order state-space model representation of an asymptotic stable LTI system:

$$\begin{pmatrix} \dot{x} \\ y \end{pmatrix} = \begin{pmatrix} A & B \\ C & D \end{pmatrix} \begin{pmatrix} x \\ u \end{pmatrix} \quad (6)$$

The problem consists in approximating the system with r^{th} order state-space model:

$$(A_r, B_r, C_r, D_r) \quad (7)$$

where $r < n$.

There are huge amount of algorithms are available in the literature for model reduction, which some of them are summarized in Figure.1.

One commonly used and globally more accurate approach is the so-called Balanced Model Reduction first introduced by Moore [7][19][20]. In this method, the system is transformed to a basis where the states which are difficult to reach are simultaneously difficult to observe. Then, the reduced model is obtained simply by truncating the states which have this property. Two other closely related model reduction techniques are Hankel Norm Approximation [11] and the Singular Perturbation method [12]. When applied to stable systems, all of these approaches are guarantees to preserve stability and provide bounds on the approximation error. Because of being operational the system within frequency band and outside that it is not important to have an accurate approximation the accuracy can be improved with applying balanced model reduction in the specified frequency band [8],[9][14-18],[22].

Controllability and observability Gramians in terms of w over a frequency band $[w_1, w_2]$ are defined as [8],[9][14-18]:



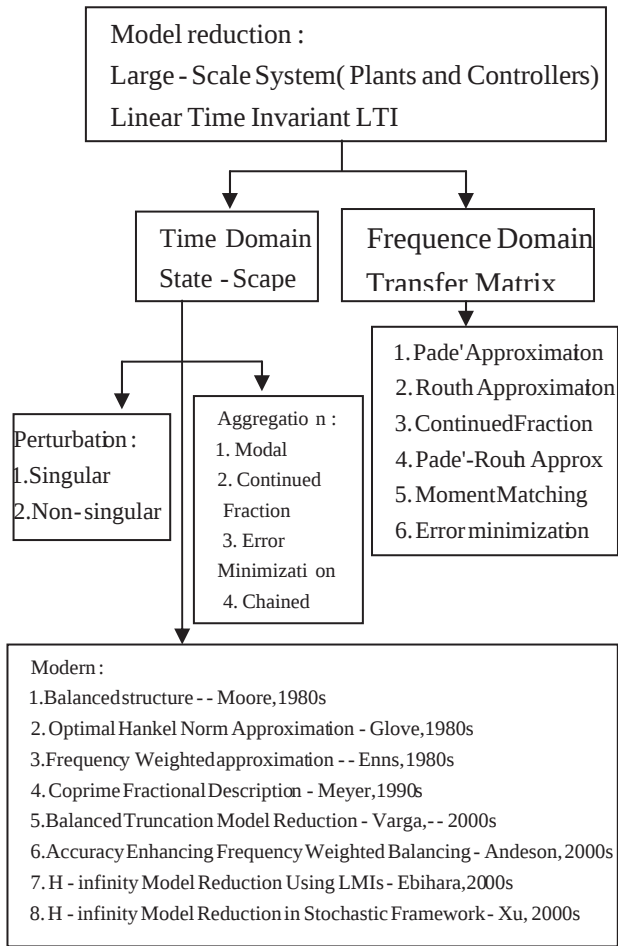


Figure.1. Algorithms for model/controller reduction.

$$W_{cf} := \frac{1}{2\pi} \int_{w_1}^{w_2} (Ij\omega - A)^{-1} B B^* (-Ij\omega - A^*)^{-1} d\omega$$

$$W_{of} := \frac{1}{2\pi} \int_{w_1}^{w_2} (-Ij\omega - A^*)^{-1} C^* C (Ij\omega - A)^{-1} d\omega$$

Those are the solutions for the following Lyapunov equations:

$$A W_{cf} + W_{cf} A^* = -(B B^* F^* + F B B^*)$$

$$A^* W_{of} + W_{of} A = -(C^* C F + F^* C^* C)$$

Where F is defined by:

$$F = \int_{w_1}^{w_2} (Ij\omega - A)^{-1} d\omega$$

With an appropriate similarity transformation T and change of the basis, system realization in (7) can be transformed to a new balanced realization, so that the Gramians are equal and diagonal (in decreasing diagonal elements):

$$\bar{W}_{cf} = \bar{W}_{of} = \text{diag}(\sigma_1, \sigma_2, \dots, \sigma_n)$$

Here we have two important theorems that give us a physical interpretation for the reduction procedure [8],[9][14-18]:

Theorem 2: The frequency-domain controllability Gramian represents the energy flow of the system through each state variable within the frequency range $[w_1, w_2]$.

It means that if the unit white Gaussian noise test input signal $u(t)$, and state vector $x(t)$ of the system defined as follows:

$$u(j\omega)[u(j\omega)]^* = |u(j\omega)|^2 = \begin{cases} 1 & w_1 < |\omega| < w_2 \\ 0 & \text{elsewhere} \end{cases}$$

$$x(t) = e^{At} B u(t) \xleftrightarrow{\mathbb{F}} (j\omega I - A)^{-1} B u(j\omega) = X(j\omega)$$

The energy of the system through controllability Gramian is as follows:

$$E_x = \int_0^\infty x(t) x^*(t) dt =$$

$$\frac{1}{2\pi} \int_{w_1}^{w_2} (j\omega I - A)^{-1} B u(j\omega) u^*(j\omega) B^* (-j\omega I - A^*)^{-1} d\omega \\ = W_{cf}[w_1, w_2]$$

Theorem 3: The frequency-domain observability Gramian represents the energy flow of the system through each state variable within the frequency range $[w_1, w_2]$.

Consider a unit injected test signal x_0 , where

$$x_0 x_0^* = \begin{cases} 1 & w_1 < \omega < w_2 \\ 0 & \text{elsewhere} \end{cases}$$

Define output

$$y(t) = C e^{At} x_0 \xleftrightarrow{\mathbb{F}} C (j\omega I - A)^{-1} x_0 = Y(j\omega)$$

The energy E_y of the system through observability Gramian is obtained by:

$$E_y = \int_0^\infty y(t) y^*(t) dt =$$

$$x_0^* \left[\frac{1}{2\pi} \int_{w_1}^{w_2} (-j\omega I - A^*)^{-1} C^* C (j\omega I - A)^{-1} d\omega \right] x_0$$

Now, x_0 being a white noise test signal, the result follows: $E_y = W_{cf}[w_1, w_2]$

From the above theorems, it is understood that for having a good approximation we should only truncate the states that are related to the lowest singular values.

3.2 Mixed method for model order reduction

In order to have more flexibility and more accuracy, the singular perturbation method and the proposed frequency balance method are combined.



The mixed method is based on the fact that every linear time-invariant system can be decomposed into the sum of two subsystems (slow and fast) in frequency domain.

System (7) can be partitioned into the slow and fast modes as:

$$\begin{pmatrix} \dot{\bar{x}}_1(t) \\ \dot{\bar{x}}_2(t) \\ y(t) \end{pmatrix} = \begin{pmatrix} \bar{A}_{11} & \bar{A}_{12} & \bar{B}_1 \\ \bar{A}_{21} & \bar{A}_{22} & \bar{B}_2 \\ \bar{C}_1 & \bar{C}_2 & D \end{pmatrix}_f \begin{pmatrix} \bar{x}_1(t) \\ \bar{x}_2(t) \\ u(t) \end{pmatrix}$$

To consider model reduction at a particular frequency level σ , we apply singular perturbation and replace $\dot{\bar{x}}_2(t)$ by $\sigma \bar{x}_2(t)$ so that σ can be chosen to be any frequency within $[w_1, w_2]$.

Setting $s = j\sigma$ and applying the method the reduced order model is obtained as:

$$\begin{pmatrix} \hat{A} \\ \hat{B} \\ \hat{C} \\ \hat{D} \end{pmatrix}_f := \begin{pmatrix} \bar{A}_{11} + \bar{A}_{12}(j\sigma I_{n-r} - \bar{A}_{22})^{-1} \bar{A}_{21} \\ \bar{B}_1 + \bar{A}_{12}(j\sigma I_{n-r} - \bar{A}_{22})^{-1} \bar{B}_2 \\ \bar{C}_1 + \bar{C}_2(j\sigma I_{n-r} - \bar{A}_{22})^{-1} \bar{A}_{21} \\ D + \bar{C}_2(j\sigma I_{n-r} - \bar{A}_{22})^{-1} \bar{B}_2 \end{pmatrix}$$

The Figure (2) shows the flowchart for the above method:

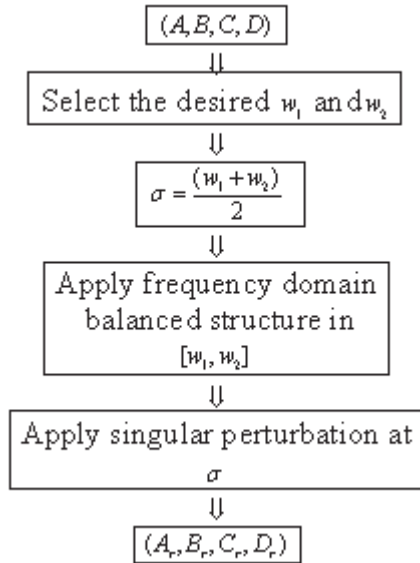


Figure 2. The mixed method Flowchart.

The idea of I/O weighting balance method is first proposed by Enns. Figure(3) shows the schematic of the I/O weight applied to a system with transfer function $G(s)$:

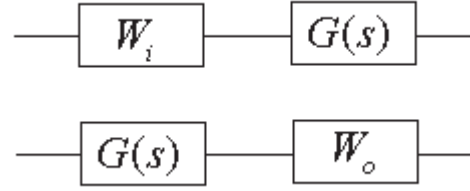


Figure 3. The schematic of I/O weights.

To combine the I/O weighting balance method and the proposed method in Figure (1) for an input or output weight W_i or W_o and the system with realization:

$$W_o := \begin{bmatrix} A_o & B_o \\ C_o & D_o \end{bmatrix}, W_i := \begin{bmatrix} A_i & B_i \\ C_i & D_i \end{bmatrix}, G(s) := \begin{bmatrix} A & B \\ C & D \end{bmatrix}$$

The overall transfer function of the structure in Figure (3) should be obtained:

$$\begin{aligned} G(s)W_i(s) &= \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} A_i & B_i \\ C_i & D_i \end{bmatrix} = \begin{bmatrix} A & BC_i & BD_i \\ O & A_i & B_i \\ C & DC_i & DD_i \end{bmatrix} \\ &= \begin{bmatrix} \tilde{A} & \tilde{B} \\ \tilde{C} & \tilde{D} \end{bmatrix} \end{aligned}$$

In the similar procedure we could obtain the overall transfer function for the output weights.

The overall transfer function obtained above is used as an input for the method proposed in Figure (2). The accuracy and performance is increased when we use the proposed mixed method and the proposed technique preserves and improves the advantages of the singular perturbation, frequency balanced and the I/O weighting reduction method.

3.3 Practical consideration

One should know how to deal with repeated or almost equal singular values in frequency-balanced realization when applying the frequency balanced truncation method. Like the usual balanced truncation. If we have two equal singular values and we only truncate the state which is related to one of them the stability of the original system may not be preserved. A large amount of the computational load of every Gramian based reduction method is imposed by finding the Gramians. In the proposed model reduction method we can easily obtain the Gramians by using the definitions and approximating the integral to summation.



3.4 Benchmark examples for frequency domain balanced reduction

In this section of the paper we apply the presented method to some benchmark examples to illustrate it more. The examples are chosen from the large scale "real word" systems reflecting current problems in applications [13].

Building Model

One of the important benchmark examples in the context of the linear time invariant systems is the building model of the Los Angeles University Hospital with 8 floors each having 3 degrees of freedom, namely displacement in x and y directions, and rotation. Hence we have 24 variables with the second order differential equation describing the dynamics of the system. The system is single input single output and the order of the system in the state-space form is 48. We apply usual balanced method and the proposed method to this model and reduce the order of it to 4. The infinity norm error of the method is shown in Figure (4). The error is lower than usual balanced method with more efficiency in computations.

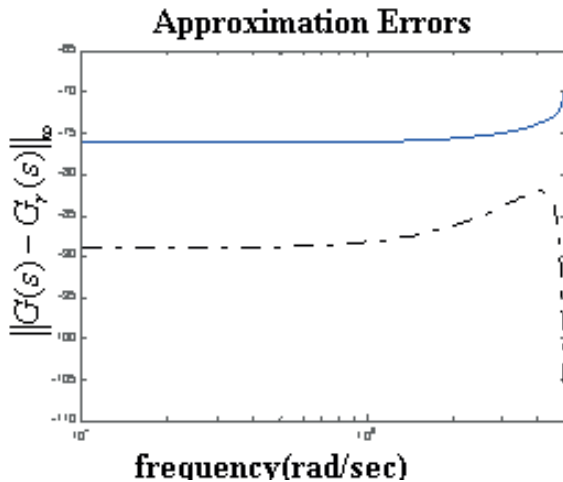


Figure 4. The approximation error in usual balanced method(solid line) and the error in the proposed method in the frequency band [0.1,4] (dash dotted).

CD Player Model

We apply the proposed method to another benchmark. This system describes the dynamics between the lens actuator and the radial arm position of a portable CD player. The model has 120 states with 2 inputs and 2 outputs. The systems is reduced to 8th order by the balanced truncation method and reduced to 4th order by the proposed method. The approximation error is shown in figure (5). Although the order of the reduced system by the proposed method is lower than the order of the

reduced system by balanced truncation technique, the reduced model by proposed balance method is more accurate than the reduced system by usual balanced truncation in some frequency ranges.

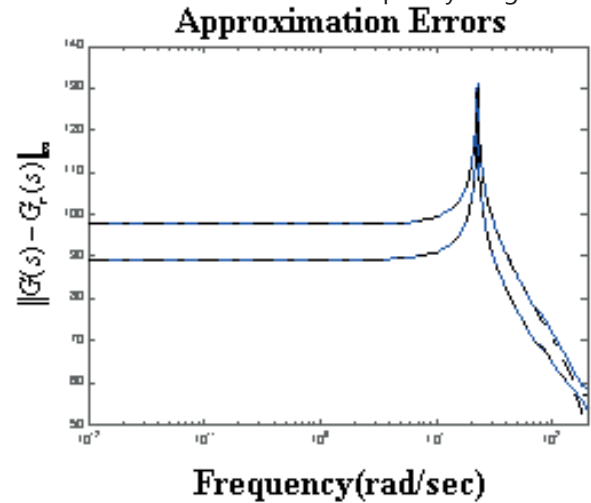


Figure 5. The approximation error in usual balanced method(solid line) and the error of the proposed method in the frequency band (dash dotted).

Atmospheric Storm Track

The model of an atmospheric storm track is a large scale single input single output model with 548 states. The system is reduced to 4th order model and the results are shown in figure (6). The infinity norm error of the proposed method is lower than the infinity error of the balanced truncation reduction method.

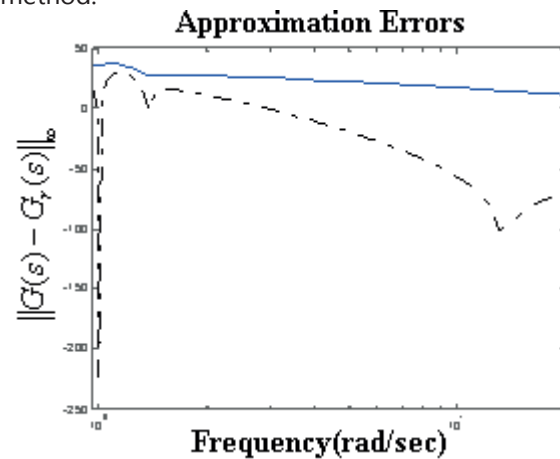


Figure 6. The approximation error in usual balanced method(solid line) and the error in the proposed method in the frequency band [0.9,20] (dash dotted).

3.5 Frequency-Domain Hybrid Reduction Method

At this point we are ready to develop new model reduction technique for reduction of large-dynamical system. First we use theorem 1 to the general structure:



$$E\dot{x}(t) = Ax(t) + Bu(t), \quad y(t) = Cx(t), \quad t \geq 0$$

and we achieve non-singular system :

$$\dot{x}_1 = (A_{11} - A_{12}A_{22}^{-1}A_{21})x_1 + (B_1 - A_{12}A_{22}^{-1}B_2)u$$

$$y = (C_1 - C_2A_{22}^{-1}A_{21})x_1 + (-C_2A_{22}^{-1}B_2)u$$

which is ready for reduction. In the first step we apply one of the methods from moment matching based reduction[6]. The second step would be applying frequency-domain balanced truncation within frequency domain which is described in detail in [10].

4. Benchmark from micro-technology: Tuneable Optical Filter

In this section we apply our method to a benchmark example in microtechnology and we compare the proposed method with the hybrid method which is based on the method in [3].

Model order reduction approach is successfully used to considerably reduce the computational time and resources. Mathematical development of MOR is an active area of research, which grows from the

reduction of linear ordinary differential equation systems (ODEs) towards the reduction of parametrized and nonlinear differential algebraic equations (DAEs) and partial differential algebraic equations (PDAEs). The implementation aspects of model order reduction are advancing as well. Practical MOR has developed from academic prototyping environments to several strong tools that can be easily used as an extension of the commercial simulators like e. g. ANSYS[10].

There are several electro-thermal MEMS models that can show the application of model reduction in Micro-Technology. We have chosen tuneable optical filter to apply our method [21].

The DFG project AFON (funded under grant ZA 276/2-1) aimed at the development of an optical filter, which is tuneable by thermal means. The thin-film filter is configured as a membrane in order to improve thermal isolation. Fabrication is based on silicon technology. Wavelength tuning is achieved through thermal modulation of resonator optical thickness, using metal resistor deposited onto the membrane. The device features low power consumption, high tuning speed and excellent optical performance.

Figure(8) shows the structure of one of the industrial benchmark examples in microtechnology. Optical tuneable filter is a micro-electromechanical system which is tuneable by thermal means. The benchmark contains a simplified thermal model of a filter device. It helps designers to consider

important thermal issues, such as what electrical power should be applied in order to reach the critical temperature at the membrane or homogeneous temperature distribution over the membrane.

The original model is the heat transfer partial differential equation. The output that we are interested in is the temperature in the center which is shown with 1 in the Figure (10). More details regarding the technical issues can be found in [10]. Figure.7 shows the procedure for using model reduction in simulation and analysis of Microsystems.

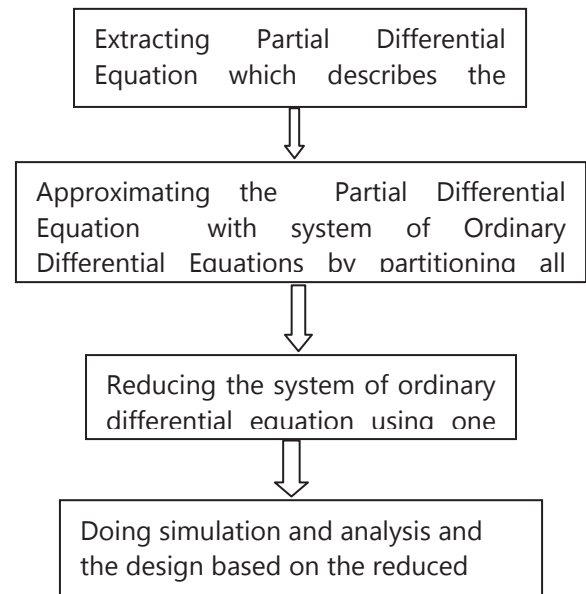


Figure 7. General Procedure for analysis and simulation for MEMS

Figure. 9 shows the procedure for using model reduction in simulation and analysis of Microsystems in our case study.

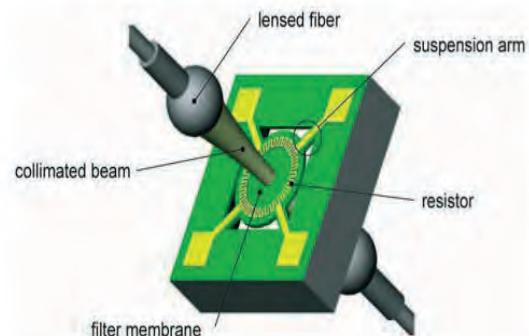


Figure 8. Tuneable optical filter

If we apply FEM(finite element method), we end up with a SISO dynamical system of order 1668 which really hard to simulate and analysis. According to our proposed framework we first reduce this dynamical system to a dynamical system of order



200 and in the next step we apply frequency-domain balanced truncation.

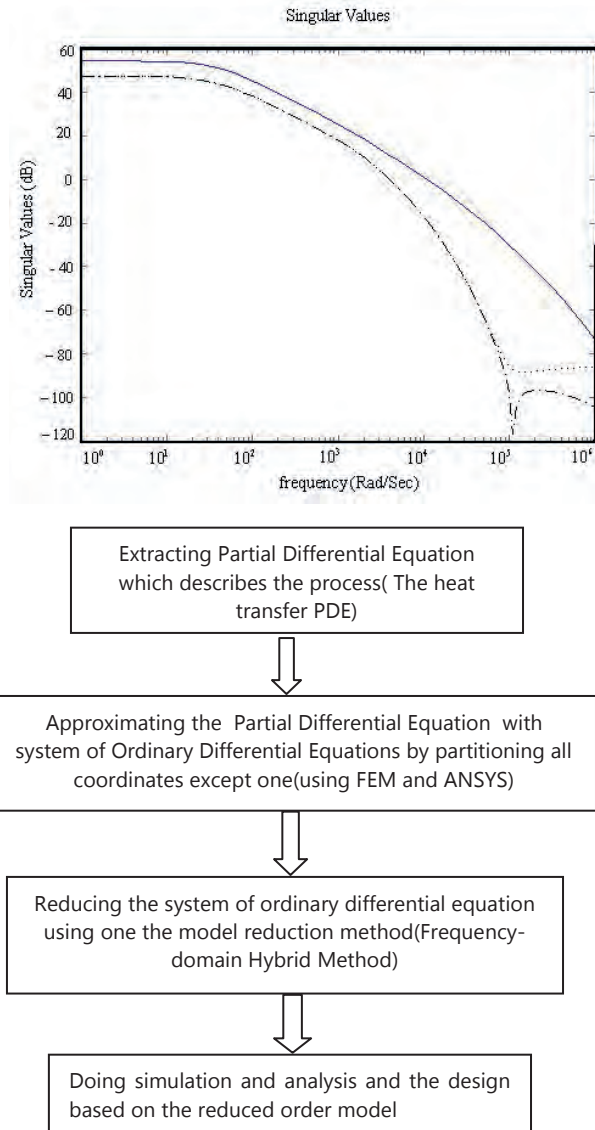


Figure 9. Procedure for analysis and simulation for MEMS in our case study

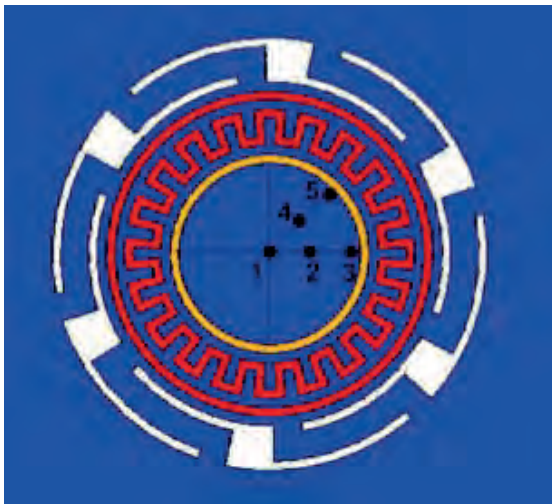


Figure 10. Cross Sectional view with outputs.

Figure 11. Singular values: original model(solid), reduced method by the proposed method(dotted) and method based on [3](dash-dotted)

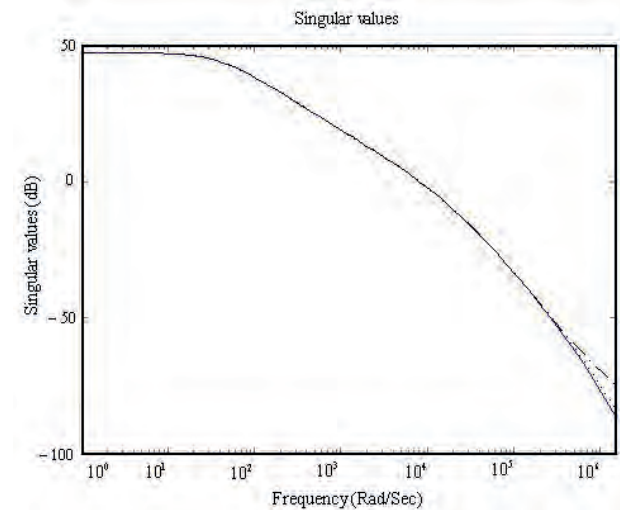


Figure 12. Singular values: original model(solid), reduced method by the proposed method(dotted) and method based on [3](dash-dotted)

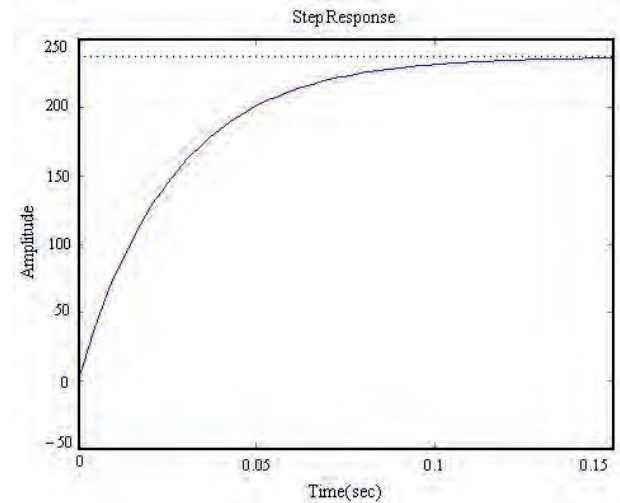


Figure 13. Step responses: original model(solid), reduced method by the proposed method(dotted) and method based on [3](dash-dotted)

Singular value of the reduced model and original systems are depicted in Figure(11) and Figure(12). In Figure (11) the frequency domain hybrid method has reduced the dynamical system to a system of order 7th and the method based on [3] to 20 and in figure (12) frequency domain hybrid method has reduced the dynamical system to a system of order 7th and the method based on [3] to 15. It is clear from the Figures that the proposed framework for reduction provides better results and can approximate the dynamical system into lower orders while preserving accuracy of the approximation. Figure(13) shows the step responses of the original and reduced models which confirms the accuracy of the approximation.



5. Conclusion

In this paper a framework for model order reduction of large dynamical systems with descriptor or singular structure is proposed. This method is a two-step framework. In the first step we apply one of the numerically efficient reduction techniques to overcome very large orders problem. Second step in the framework is frequency-domain balanced reduction within frequency bound. This framework provides more efficient and accurate results comparing to the previous counterparts. This reduction technique is developed for model reduction of large dynamical singular systems which are frequently appear in analysis and simulation in Micro-technology as a result of finite element method. Numerical results show the accuracy and efficiency enhancement of the method in comparison to previous techniques in this context.

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Hybrid PWM Algorithm with Low Computational Overhead for Induction Motor Drives for Reduced Current Ripple

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Abstract

This paper presents an analytical approach for the analysis of the current ripple in space vector based PWM controlled induction motor drives. This paper considers seven different types of switching sequences. By using the instantaneous ripple voltages, the current ripple expressions are derived for all the sequences. To reduce the complexity involved in the conventional space vector approach, the proposed sequences have been carried out using the concept of imaginary switching times. Finally, to reduce the current ripple, hybrid PWM (HPWM) algorithm has been developed which uses a sequence that results in the lowest rms current ripple over the given sampling time interval. Thus, the proposed HPWM algorithm reduces the current ripple at all modulation indices. To validate the proposed HPWM algorithm, numerical simulation studies have been carried out and compared with the conventional space vector PWM (CSVPWM) algorithm.

Keywords: current ripple, hybrid PWM, induction motor, SVPWM.

1. Nomenclature

V_{an}, V_{bn}, V_{cn} : Instantaneous phase voltages

T_{an}, T_{bn}, T_{cn} : Imaginary switching times

T_1, T_2 : Active vectors switching times

T_z : Zero voltage vectors switching time

$T_{max}, T_{min}, T_{mid}$: Maximum, minimum and middle values of imaginary switching times

V_{rip} : Ripple voltage vector

V_{ref} : Reference voltage vector

α : Angle of the reference voltage vector

i_{rip} : Current ripple vector

i_{rms} : Rms value of current

M_i : Modulation index

σ : Leakage coefficient of induction motor

L_s : Stator inductance in H

2. Introduction

Recently, the voltage source inverters are becoming popular to feed variable voltage, variable frequency voltages to the three-phase induction motor drives. A detailed survey on the various PWM algorithms for voltage source inverters is given in [1]. Among the various PWM algorithms, SVPWM algorithm offers many advantages and hence it is becoming popular [2]. Moreover, among the various approaches, the triangular comparison (TC) approach and space vector (SV) approach are popular. However, the SV approach offers so many degrees of freedom for the implementation [3]. However, the conventional SVPWM (CSVPWM) algorithm uses equal division of zero state time duration. By utilizing the freedom of active state division, nowadays many PWM algorithms are derived [3]. By using only one zero state, many bus-clamping PWM algorithms can be generated [4]. But the existing bus-clamping PWM algorithms, which are given in [4] did not involve in the active state division. Similar to the zero state time division, the active state time division also can be performed in the SV approach. Recently, these are attracting the researchers and a detailed study about these algorithms are given [3], [5]-[7], [9], [11]. The double



switching sequences (involving active state division) give superior performance at higher modulation indices. To reduce the steady state ripples at all modulation indices hybrid PWM (HPWM) algorithms are developed by using the notion of stator flux ripple [5]-[7], [9], [11].

However, the conventional SV approach requires the calculation of angle and sector information in every sampling time interval and hence increases the complexity. To reduce the complexity involved in the conventional SV approach, a novel approach has been developed by utilizing the concept of offset time in [10]. To reduce the burden involved in conventional HPWM algorithms, novel HPWM algorithm using the concept of imaginary switching times is given in [11]. However, the proposed HPWM algorithms in [5]-[7], [11] uses the notion of stator flux ripple to develop the HPWM algorithm. To reduce the current ripple, the analysis has been carried out using the notion of the current ripple in [8]-[9].

This paper presents a novel HPWM algorithm for reduced current ripple which uses the notion of stator current ripple and uses the concept of imaginary switching times [10]-[11].

The remainder of the paper is organized as follows: Section (2) focuses on the generation of various switching sequences using the concept of imaginary switching times. Section (3) emphasizes on the analysis of stator current ripple. Section (4) focuses on the development of HPWM algorithms and section (5) gives the simulation results and discussions and conclusions are given in section (6).

3. Switching Sequences

The conventional SV approach uses reference frame transformation and angle calculation. Hence, it increases the complexity involved in the algorithm. But, the proposed approach uses the instantaneous reference phase voltages for the generation of switching sequences. For the explanation purpose, the concept of imaginary switching times will be introduced. The imaginary switching times are proportional to the instantaneous phase voltages and can be defined as follows:

$$T_{an} = \frac{T_s}{V_{dc}} V_{an}; T_{bn} = \frac{T_s}{V_{dc}} V_{bn}; T_{cn} = \frac{T_s}{V_{dc}} V_{cn} \quad (1)$$

V_{an} , V_{bn} and V_{cn} are the instantaneous reference phase voltages and T_{an} , T_{bn} and T_{cn} are the corresponding imaginary switching times. As these

times are proportional to the instantaneous voltages, these times could be negative where the voltages are negative. Hence, these times are defined as imaginary switching times. The active vector switching times T_1 and T_2 , if the reference voltage vector falls in sector-1 may be expressed as follows [11]:

$$T_1 = T_{an} - T_{bn}; T_2 = T_{bn} - T_{cn} \quad (2)$$

In the conventional SVPWM algorithm, when the reference voltage vector falls in the first sector, the imaginary switching time which is proportional to the a-phase (T_{an}) has a maximum value, the imaginary switching time which is proportional to the c-phase (T_{cn}) has a minimum value and the imaginary switching time which is proportional to the b-phase (T_{bn}) is neither minimum nor maximum value. Thus, in general to calculate the active vector switching times, the maximum (T_{max}), middle (T_{mid}) and minimum (T_{min}) values of imaginary switching times are calculated in every sampling time. Then the active vector switching times T_1 and T_2 may be expressed as

$$T_1 = T_{max} - T_{mid}; T_2 = T_{mid} - T_{min} \quad (3)$$

The maximum and minimum imaginary switching times and active vector switching times for all six sectors are given in Table 1.

The zero voltage vectors switching time is calculated as

$$T_z = T_s - T_1 - T_2 \quad (4)$$

To generate the various switching sequences, the zero state time will be shared between two zero states as xT_z for V_0 and $(1-x)T_z$ for V_7 respectively.

Thus, by using (3) and (4), the active vector times and zero vector times can be calculated. If $x=0.5$, then the conventional SVPWM algorithm is obtained. When $x=0$, any one of the phases is clamped to positive dc bus for 120 degrees over a fundamental cycle. When $x=1$, any one of the phases is clamped to negative dc bus for 120 degrees over a fundamental cycle. Then by utilizing the space vector concept and zero and active vector division concept, various space vector based PWM algorithms can be generated. The existing bus-clamping (BCPWM) algorithm use only one zero state and the advanced BCPWM (ABCPWM) sequences use only one zero state and active state division. The sequences of all the possible PWM algorithms in six sectors are listed in Table-2.



Table 1: Various times in all six sectors

	Sector-I	Sector-II	Sector-III	Sector-IV	Sector-V	Sector-VI
$T_{\max}$	T_{an}	T_{bn}	T_{bn}	T_{cn}	T_{cn}	T_{an}
$T_{\min}$	T_{cn}	T_{cn}	T_{an}	T_{an}	T_{bn}	T_{bn}
T_1	$T_{an} - T_{bn}$	$T_{bn} - T_{an}$	$T_{bn} - T_{cn}$	$T_{cn} - T_{bn}$	$T_{cn} - T_{an}$	$T_{an} - T_{cn}$
T_2	$T_{bn} - T_{cn}$	$T_{an} - T_{cn}$	$T_{cn} - T_{an}$	$T_{bn} - T_{an}$	$T_{an} - T_{bn}$	$T_{cn} - T_{bn}$

Table 2: Sequences of various possible PWM algorithms

Sector	CSVPWM	BCPWM algorithms		ABCPWM algorithms			
I	0127-7210	012-210	721-127	0121-1210	7212-2127	1012-2101	2721-1272
II	0327-7230	032-230	723-327	0323-3230	7232-2327	3032-2303	2723-3272
III	0347-7430	034-430	743-347	0343-3430	7434-4347	3034-4303	4743-3474
IV	0547-7450	054-450	745-547	0545-5450	7454-4547	5054-4505	4745-5474
V	0567-7650	056-650	765-567	0565-5650	7656-6567	5056-6505	6765-5676
VI	0167-7610	016-610	761-167	0161-1610	7616-6167	1016-6101	6761-1676

4. Analysis of the Current Ripple

The space vector approach generates the reference voltage vector in an average manner. But, when the required voltage vector is applied to the motor, the current is flowing in the stator winding of the induction motor. The actual value of the stator voltage differs from the applied voltage. Hence, there is always an instantaneous error voltage vector. The error voltage vector is defined as given in (5)

$$V_{rip} = V_k - V_{ref} \quad (5)$$

where 'k' is the k^{th} voltage vector and it takes any value from '0' to '7'. Then, it is possible to define the current ripple vector as defined in (6)

$$i_{rip} = \frac{1}{\sigma L_s} \int_0^t V_{rip} dt \quad (6)$$

The voltage ripple vectors and the variation of the current ripple by the application of corresponding voltage vectors are shown in figure 1 for all the possible PWM algorithms. Also, from this figure, it can be observed that the current ripple has zero mean value in a sampling time period. Then the current ripple vectors are given by as follows:

$$i_{r1} = \frac{V_1 - V_{ref}}{\sigma L_s} T_1 \quad (7)$$

$$i_{r2} = \frac{V_2 - V_{ref}}{\sigma L_s} T_2 \quad (8)$$

$$i_{rz} = -\frac{V_{ref}}{\sigma L_s} T_z \quad (9)$$

The current ripple vectors can be represented in terms of imaginary switching times as given in (10)-(12).

$$i_{r1} = i_d + ji_{q1} \quad (10)$$

$$i_{r2} = i_d + ji_{q2} \quad (11)$$

$$i_{rz} = -j \frac{T_z}{\sigma L_s} \frac{2M_i V_{dc}}{\pi} = -ji_{qz} \quad (12)$$

where M_i is the modulation index and defined as

$$M_i = \frac{\pi V_{ref}}{2V_{dc}} \quad (13)$$

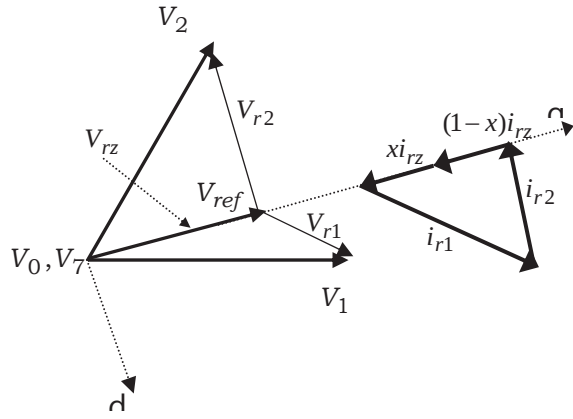
$$i_d = \frac{T_1}{\sigma L_s} \left(\frac{\pi V_{dc}}{3\sqrt{3}M_i} \frac{T_2}{T_s} \right) \quad (14)$$

$$i_{q1} = \frac{T_1}{\sigma L_s} \left(\frac{2V_{dc}\pi(T_1 + 0.5T_2)}{9M_i T_s} - \frac{2V_{dc}M_i}{\pi} \right) \quad (15)$$

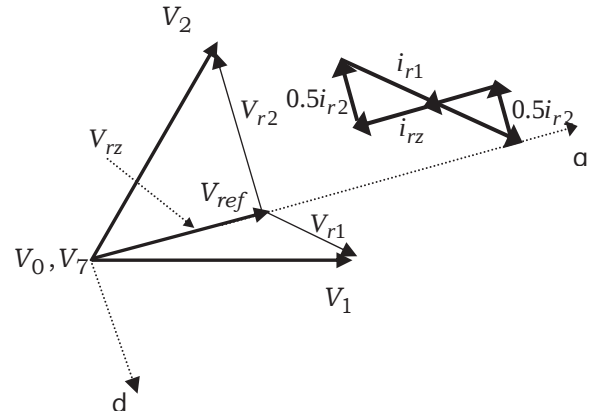
$$i_{q2} = \frac{T_2}{\sigma L_s} \left(\frac{2V_{dc}\pi(0.5T_1 + T_2)}{9M_i T_s} - \frac{2V_{dc}M_i}{\pi} \right) \quad (16)$$

Then the current ripple vectors can be resolved along the d - and q -axes, which are the synchronously revolving reference frame axes as shown in figure 2, from which it can be observed that the application of a zero state results in a variation of the q -axis component of the current ripple and the application of any active state results in variation of the both the d -axis and q -axis current components.



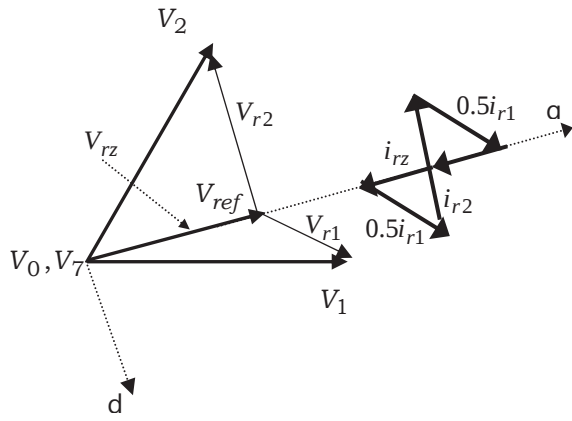


(a)

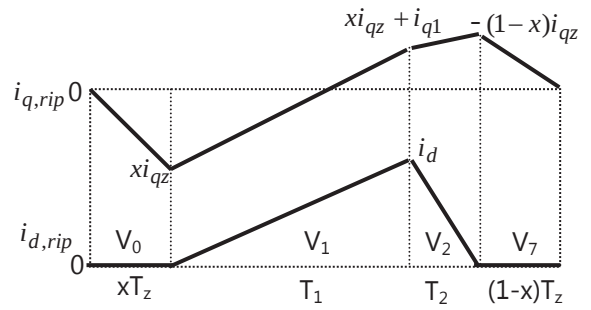


(e)

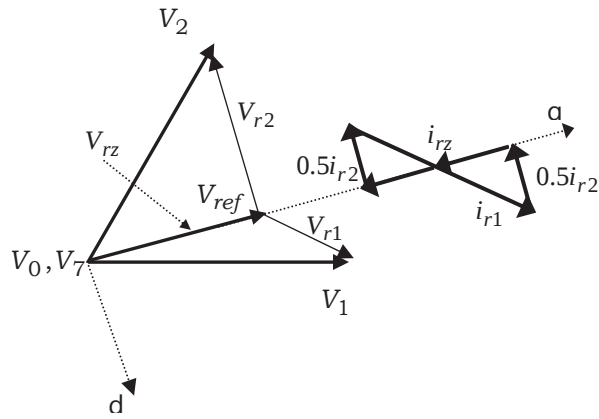
Figure 1: Error voltage vectors along with the current ripple variations (a) 0127, 012, 721 sequences (b) 0121 (c) 7212 (d) 1012 (e) 2721



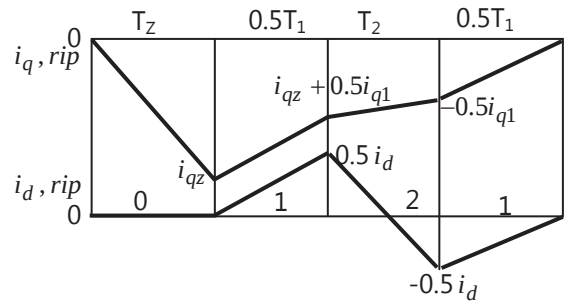
(b)



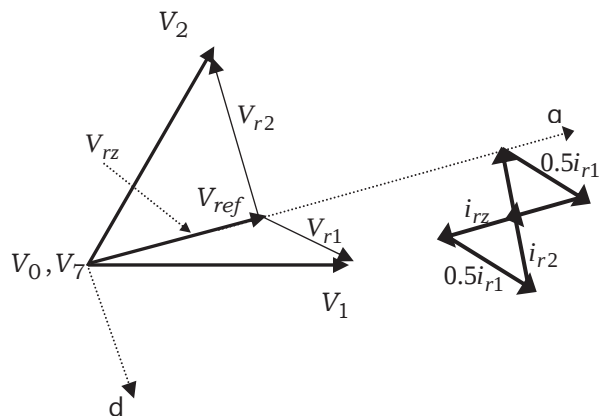
(a)



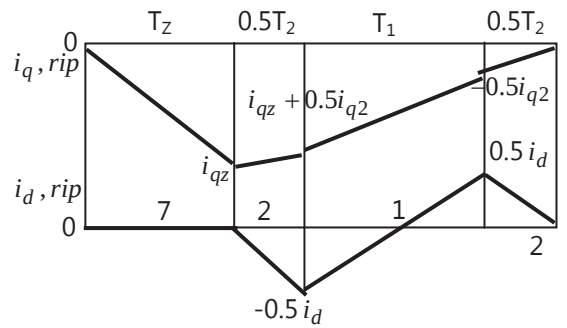
(c)



(b)



(d)



(c)



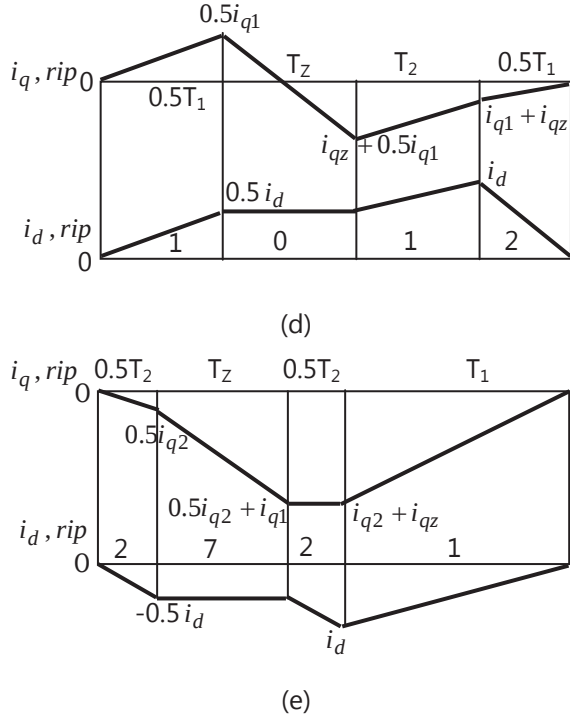


Figure 2: the q-and d-axes components of the current ripple vectors for the possible sequences. (a) 0127, 012, 721 sequences (b) 0121 (c) 7212 (d) 1012 (e) 2721

The q-axis, d-axis and the total rms current ripple over a sampling time interval (T_s) can be evaluated as follows:

$$i_{q,rms}^2 = \frac{1}{T_s} \int_0^{T_s} i_{q,rip}^2 dt \quad (17)$$

$$i_{d,rms}^2 = \frac{1}{T_s} \int_0^{T_s} i_{d,rip}^2 dt \quad (18)$$

$$i_{rms}^2 = i_{q,rms}^2 + i_{d,rms}^2 \quad (19)$$

Finally, the rms current ripple expressions can be as given in (20) – (24).

$$i_{rms}^2 = \frac{1}{3} \left\{ x i_{q0}^2 \frac{x T_z}{T_s} + \left[x i_{qz}^2 + (i_{qz} + i_{q1})^2 + i_{qz}(i_{qz} + i_{q1}) \right] \frac{T_1}{T_s} + \left[(i_{qz} + i_{q1})^2 - (1-x) i_{qz}(i_{qz} + i_{q1}) + ((1-x) i_{qz})^2 \right] \frac{T_2}{T_s} + \left[((1-x) i_{qz})^2 \frac{T_2}{T_s} + i_d^2 \frac{(T_1 + T_2)}{T_s} \right] \right\} \quad (20)$$

$$i_{rms0121}^2 = \frac{1}{3} \left\{ i_{qz}^2 \frac{T_z}{T_s} + \left[i_{qz}^2 + i_{qz}(i_{qz} + 0.5i_{q1}) + (i_{qz} + 0.5i_{q1})^2 \right] \frac{T_1}{2T_s} + \left[(i_{qz} + 0.5i_{q1})^2 - (i_{qz} + 0.5i_{q1})0.5i_{q1} + (0.5i_{q1})^2 \right] \frac{T_2}{T_s} + \left[(0.5i_{q1})^2 \frac{T_1}{2T_s} + (0.5i_d)^2 \frac{(T_1 + T_2)}{T_s} \right] \right\} \quad (21)$$

$$i_{rms7212}^2 = \frac{1}{3} \left\{ i_{qz}^2 \frac{T_z}{T_s} + \left[i_{qz}^2 + i_{qz}(i_{qz} + 0.5i_{q2}) + (i_{qz} + 0.5i_{q2})^2 \right] \frac{T_2}{2T_s} + \left[(i_{qz} + 0.5i_{q2})^2 - (i_{qz} + 0.5i_{q2})0.5i_{q2} + (0.5i_{q2})^2 \right] \frac{T_1}{T_s} + \left[(0.5i_{q2})^2 \frac{T_2}{2T_s} + (0.5i_d)^2 \frac{(T_1 + T_2)}{T_s} \right] \right\} \quad (22)$$

$$i_{rms1012}^2 = \frac{1}{3} \left\{ (0.5i_{q1})^2 \frac{T_1}{2T_s} + \left[(0.5i_{q1})^2 + 0.5i_{q1}(0.5i_{q1} + i_{qz}) + (0.5i_{q1} + i_{qz})^2 \right] \frac{T_z}{T_s} + \left[(0.5i_{q1} + i_{qz})^2 + (0.5i_{q1} + i_{qz})(i_{q1} + i_{qz}) + (i_{q1} + i_{qz})^2 \right] \frac{T_1}{2T_s} + \left[(i_{q1} + i_{qz})^2 \frac{T_2}{T_s} + i_d^2 \frac{(T_1 + T_2)}{T_s} + (0.5i_d)^2 \frac{T_z}{T_s} \right] \right\} \quad (23)$$

$$i_{rms2721}^2 = \frac{1}{3} \left\{ (0.5i_{q2})^2 \frac{T_2}{2T_s} + \left[(0.5i_{q2})^2 + 0.5i_{q2}(0.5i_{q2} + i_{qz}) + (0.5i_{q2} + i_{qz})^2 \right] \frac{T_z}{T_s} + \left[(0.5i_{q2} + i_{qz})^2 + (0.5i_{q2} + i_{qz})(i_{q2} + i_{qz}) + (i_{q2} + i_{qz})^2 \right] \frac{T_2}{2T_s} + \left[(i_{q2} + i_{qz})^2 \frac{T_1}{T_s} + i_d^2 \frac{(T_1 + T_2)}{T_s} + (0.5i_d)^2 \frac{T_z}{T_s} \right] \right\} \quad (24)$$

The number of switchings of the CSVPWM and ABCPWM sequences in a sampling time interval is 3 and whereas for the BCPWM algorithms is 2. Hence, to get the constant average switching frequency of the inverter, a sampling time interval is taken as $T_s = T$ for the CSVPWM and ABCPWM algorithms, while $T_s = (2T/3)$ for the BCPWM algorithms. Thus the proposed method of analysis is capable of estimating d-axis and q-axis stator current ripples individually. Total harmonic distortion (THD) of the current waveform is equally affected by the d-axis and q-axis stator current ripples. On the other hand, torque pulsation mainly depends on q-axis current ripple, and is practically independent of d-axis stator current ripple. Thus, the reduction in q-axis current ripple signifies the reduction of THD and torque



pulsation, while the reduction in d-axis current ripple implies reduction in THD only. By using (20)-(23), the rms current ripple trajectories over a sixty degrees period for every PWM sequence at different modulation indices are plotted as shown in figure 3.

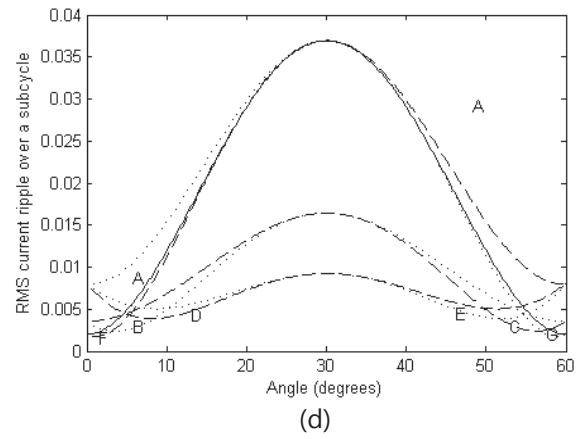
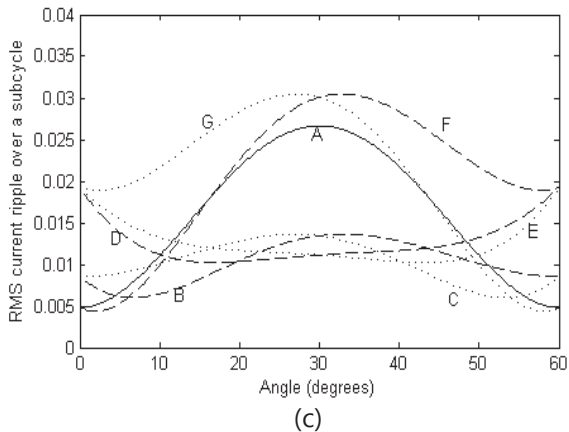
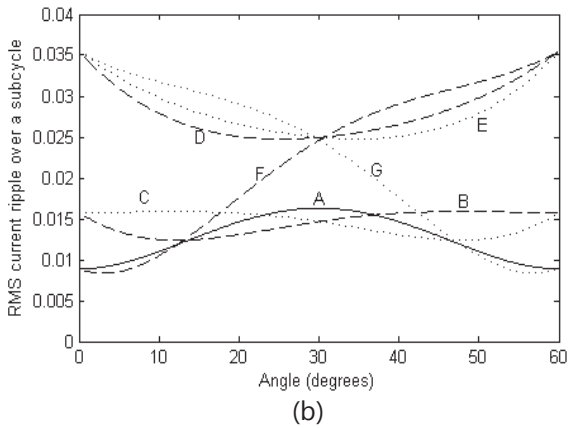
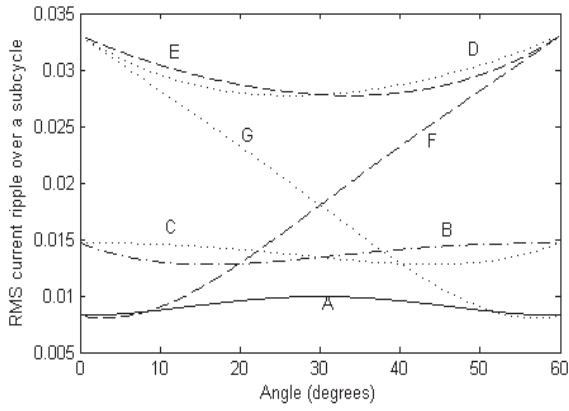


Figure 3: current ripple trajectories over a sixty degrees period (for one sector) (a) $M_i=0.4$, (b) $M_i=0.6$, (c) $M_i=0.8$ and (d) $M_i=0.905$ (A=0127, B=012, C=721, D=0121, E=7212, F=1012 and G= 2721)

5. Development of HPWM Algorithms

From figure 3, it can be observed that replacing of α by $(60^\circ - \alpha)$ in the current ripple expression of CSVPWM sequence does not change their values. Thus for a given sampling time and M_i , 0127 sequence produces equal mean square current ripple at spatial angles α and $(60^\circ - \alpha)$. Whereas for the other sequences the mean square current ripple at α is equal to the mean square stator current ripple of their respective complimentary sequences at $(60^\circ - \alpha)$. For a given values of M_i and T_s , these can be given as follows:

$$i_{rms,012}^2(\alpha) = i_{rms,721}^2(60^\circ - \alpha) \quad (25)$$

$$i_{rms,0121}^2(\alpha) = i_{rms,7212}^2(60^\circ - \alpha) \quad (26)$$

$$i_{rms,1012}^2(\alpha) = i_{rms,2721}^2(60^\circ - \alpha) \quad (27)$$

By comparing the sequences 0127, 012, 721, 0121, 7212, 1012 and 2721 with respect to each other, the zones of superior performance of each sequence can be identified. The development of HPWM techniques for reduced current ripple involves determination of superior performance for every sequence. The zone of superior performance for a given sequence is the spatial zone within a sector where the given sequence results in less mean square current ripple than other sequences considered. Thus, the proposed HPWM algorithm is a 7-zone PWM algorithm, which uses six bus-clamping PWM algorithms along with the CSVPWM algorithm for reduced steady state current ripple.

The flowchart of proposed HPWM algorithm is as shown in figure 4.



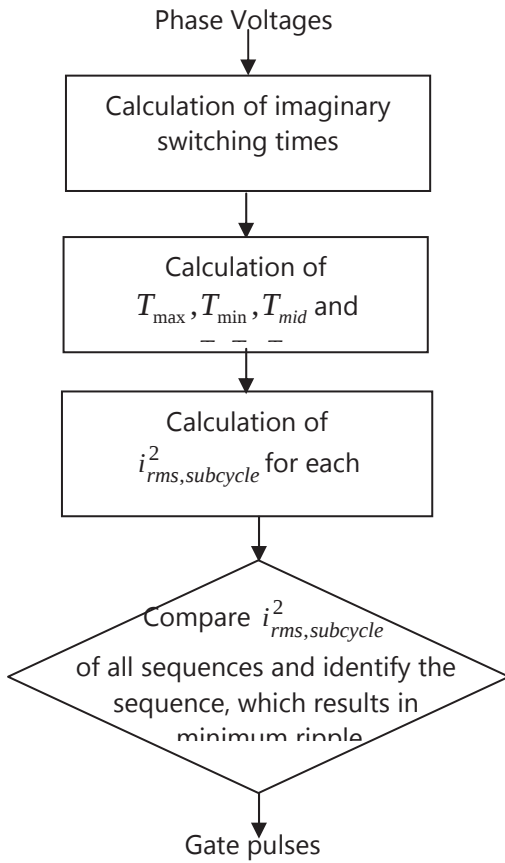


Figure 4: Flowchart of proposed HPWM algorithm

6. Simulation Results and Discussion

To verify the proposed HPWM algorithm, numerical simulation studies have been carried out on a v/f controlled induction motor drive by using Matlab/Simulink. For the simulation, the average switching frequency of the inverter is taken as 5 kHz. The induction motor used in this case study is a 4 KW, 400V, 1470 rpm, 4-pole, 50 Hz, 3-phase induction motor having the following parameters:

$R_s = 1.57\Omega$, $R_r = 1.21\Omega$, $L_s = 0.17H$, $L_r = 0.17H$, $L_m = 0.165 H$ and $J = 0.089 \text{ Kg.m}^2$. The steady state currents along with their harmonic spectra are given in Figure.5 – Figure.8 for SVPWM and proposed HPWM algorithms based v/f controlled induction motor drives at different supply frequencies. From the simulation results, it can be observed that the proposed HPWM algorithm reduces the total harmonic distortion (THD) in the line current when compared with the CSVPWM algorithm. Moreover, from the line current spectra corresponding to CSVPWM algorithm, it can be observed that the harmonic components are around frequencies that are integral multiples of the switching frequency (5 kHz). On the other hand, the proposed HPWM algorithm gives spread spectra. Hence, the proposed HPWM algorithm also reduces the acoustical noise of the induction motor with a

uniform sampling frequency. Thus, the proposed HPWM algorithm reduces the steady state current ripple and acoustical noise of the motor when compared with the CSVPWM algorithm.

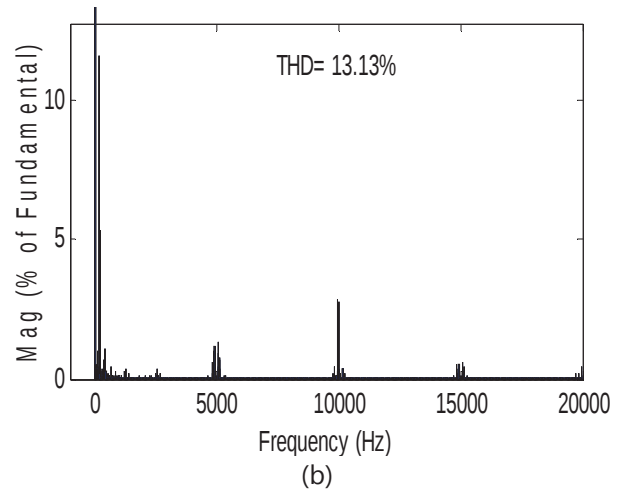
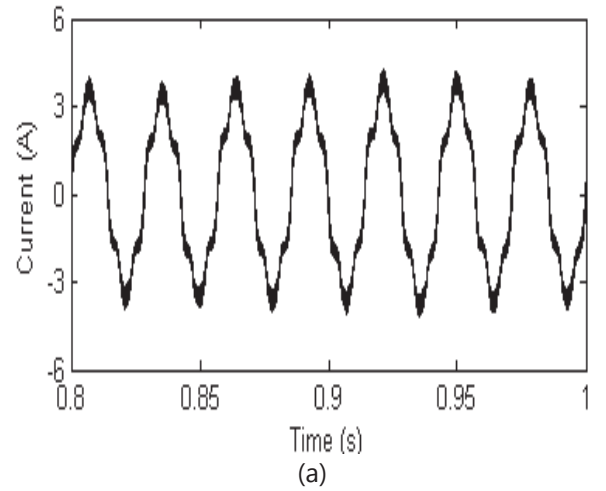
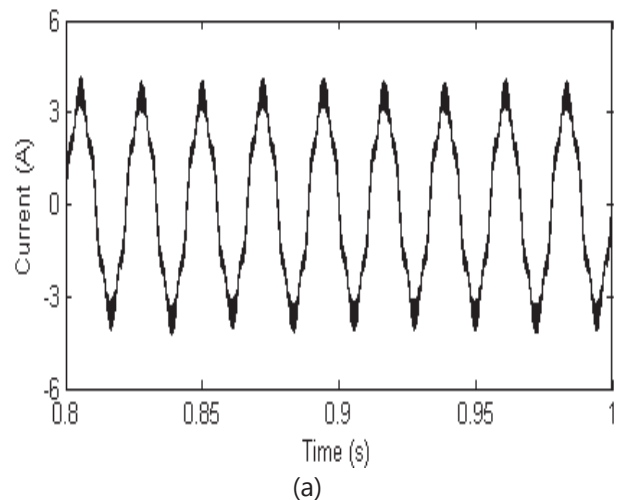


Figure 5: Steady state results with SVPWM algorithm at a supply frequency of 35 Hz, (a) Steady state current (b) current spectra



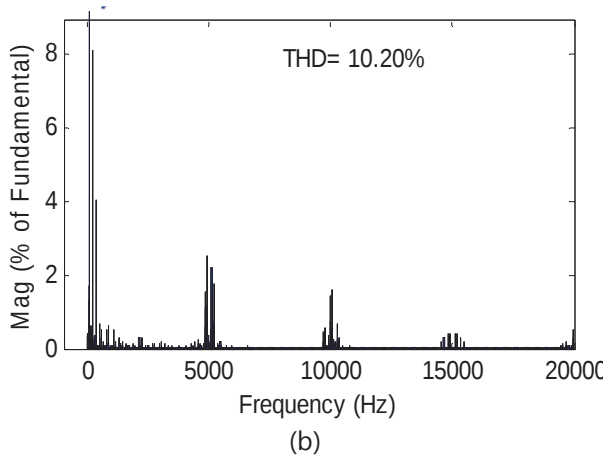


Figure 6: Steady state results with SVPWM algorithm at a supply frequency of 45 Hz, (a) Steady state current (b) current spectra

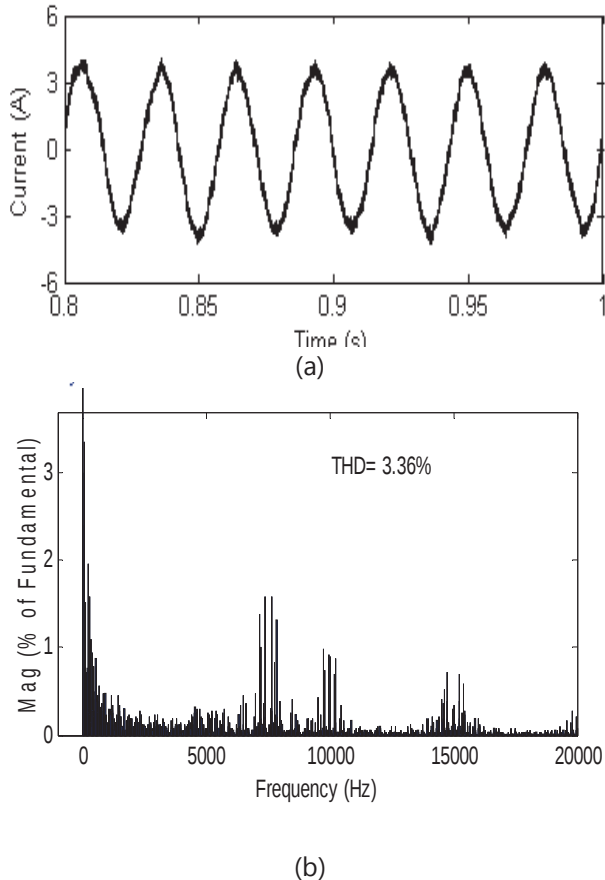


Figure 7: Steady state results with HPWM algorithm at a supply frequency of 35 Hz, (a) Steady state current (b) current spectra

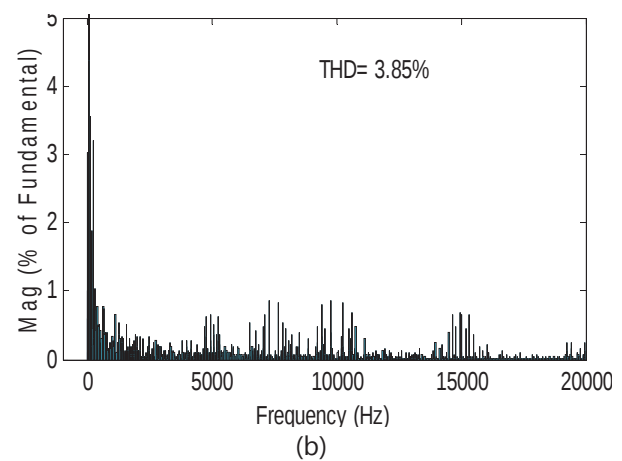
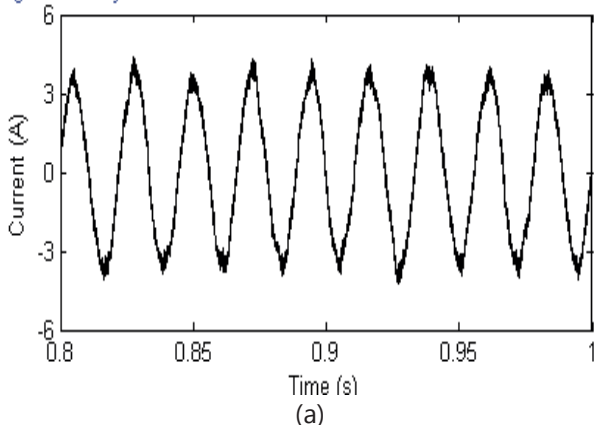


Figure 8: Steady state results with HPWM algorithm at a supply frequency of 45 Hz, (a) Steady state current (b) current spectra

7. Conclusions

A simplified HPWM algorithm is presented in this paper for reduced current ripple. This paper has analyzed the instantaneous current ripple and derived the expressions for the variation of current ripple for the possible PWM algorithms. Finally, HPWM algorithm has developed in this paper. From the simulation results, it can be observed that the proposed HPWM algorithm gives less THD when compared with the conventional SVPWM algorithm over a wide range of operating frequencies (modulation indices). Moreover, as the HPWM gives spread spectra, it reduces the acoustical noise of the induction motor with a uniform sampling frequency. Hence, to reduce the acoustical noise, special methods with random sampling frequencies are not necessary. Thus, the proposed HPWM algorithm gives better performance at all modulation indices.

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Fuzzy Logic Based Control for Parallel Cascade Control

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Abstract

In chemical industries parallel cascade controllers are used to improve the dynamic performance of a control system. Cascade controllers has two controllers, primary and secondary controllers in cascade. In this paper, a fuzzy based control is designed for the primary controller to improve the performance of the parallel cascade control. The secondary controller is implemented with internal model control (IMC) method. The dead time in the process is compensated by incorporating Smith predictor in the primary loop. The use of fuzzy logic controller is more effective than the conventional controller. It contributes to achieve quick response and high accuracy to variations in input parameter fluctuations during operations in systems, and also provides robust control performance. The performance of the proposed method has been evaluated by conducting experiments on two different systems. The simulation results revealed that the proposed method outperforms other recently reported methods in performance measures for both transient and steady state conditions.

Keywords: Parallel cascade control, PID control, Fuzzy logic control, Dead time, Internal model control

1. Nomenclature

G_{p1}, G_{p2} Primary and secondary processes transfer function

G_{d1}, G_{d2}	Transfer functions of disturbances (d)
G_{c1}, G_{c2}	Primary and secondary controller transfer function
d	Disturbance input for primary and secondary loop
r_1, r_2	Primary and secondary loop set point
Y_1, Y_2	Primary and secondary process output
G_{p2m}	Secondary process model
G_{f2}	Set point filter
G_{f1}	Filter for predicted disturbance
G_{pm}	Process model of the outer loop without dead time
θ_m	Dead time of the model of the outer loop
u	Crisp valued controller output
f_i	Fuzzy valued controller output
μ_i	Input membership function

2. Introduction

In conventional single feedback control, the corrective action for disturbance does not begin until the controlled variable deviates from the set point. If the response is unsatisfactory because of large uncontrolled load changes, more complex control schemes can be considered [1]. Cascade control, first introduced by Franks and Worley [2], is one of the frequently used schemes to deploy the availability of additional controllers for the plant to



effectively adopt to wide fluctuations in loads. The concept of cascade control is to nest one feedback loop inside another loop. Cascade control involves the use of a primary and secondary controller. In process industries, cascade control is widely used to reduce the effects of possible disturbances and to improve the dynamic performance of the cascade loop system. In traditional series cascade control, both the manipulated variable and the disturbance affect the primary output through the intermediate (secondary) output. The cascade control of parallel type arises when the manipulated and disturbance variables simultaneously affect primary and secondary outputs [3]. The overhead composition control of a distillation column, cascaded onto the control of a tray temperature, is a typical example of a parallel cascade control. The reflux flow rate (manipulated variable) and the feed flow or composition (disturbance) affect both, the purity of the overhead product (primary output) and the tray temperature (secondary output). The control objective is to maintain the overhead composition at set point. The output of the composition controller resets the set point for the temperature controller. By controlling the tray temperature in the cascade manner, the variation in the feed can be compensated before it disturbs the product composition. Cascade control is used to improve the dynamic response of the column [4]. In general, parallel cascade control is appropriate when the secondary loop has a faster dynamic response and the rejection of the disturbance in the secondary output reduces the steady state output error in the primary loop. The parallel cascade control is also beneficial when measurements of the primary output are sampled infrequently and/or with long time delays.

Though parallel cascade control is widely used in industries, it has attracted very little research despite its clear benefits. Yu [5] proposed an interaction measure to determine whether the parallel cascade control is advantageous or detrimental to load response. Shen and Yu [6] applied these results to the selection of secondary measurement in parallel cascade control. Bramilla and Semino [7] introduced a nonlinear filter in order to practically separate the dynamics of the primary and secondary loop in

parallel cascade control. They also developed a combined structure to avoid interactions between primary and secondary loops [8]. Mcavoy and Ye [9] discussed nonlinear inferential parallel cascade control structure based on both parallel cascade control and inferential sensing. Pottman et al. [10] developed a parallel control strategy for a biological control system. Chen et al. [11] focused on the performance assessment of parallel cascade control systems using the methodology of univariate control loop performance, minimum variance and Diophantine decomposition principles. Lee et al. [12] proposed an analytical method of Proportional Integral Derivative (PID) controller design for parallel cascade control system which takes into account the interaction between primary and secondary loops. Recently, Seshagiri et al. [13] incorporated time delay compensator in the existing parallel cascade control. Most industrial processes and the primary loop in cascade control systems in particular, are non-linear, multivariable, distributed complex dynamic systems. There are also large delays in dynamic channels and cross relationship between control parameters. The process dead time and gain are non-linear and are not predictable because they vary according to process conditions [14]. For instance, changes in flow, pressure, temperature, or level may cause large variations in process dead time and gain. This makes some process control loops extremely non-linear. Due to interconnections, dead time and non-linearity, the control of these systems become difficult [15]. Also, the accurate process models are not known. When there is a lack of accurate process model, the existing controllers designed so far will not be much effective. Fuzzy control strategy is fit for systems that lacks accurate model. In this paper, fuzzy control is designed for the primary controller. It is a well known method of implementing nonlinear control. Hence the parallel cascade control is improved by incorporating fuzzy controller in controller design.

This paper is organized as follows: Next section gives brief overview of parallel cascade control. Section 4 provides the secondary and primary controller design procedure and simulation result for the different case studies to illustrate the use of



proposed cascade control. Conclusions are provided in section 5.

3. Parallel Cascade Control

The parallel cascade control structure is shown in Figure 1. G_{p1} and G_{p2} are the transfer functions of the

primary and secondary processes respectively. G_{d1} and G_{d2} are the transfer functions of disturbances (d) for the primary and secondary loops. G_{c1} and G_{c2} are the primary and secondary controllers respectively.

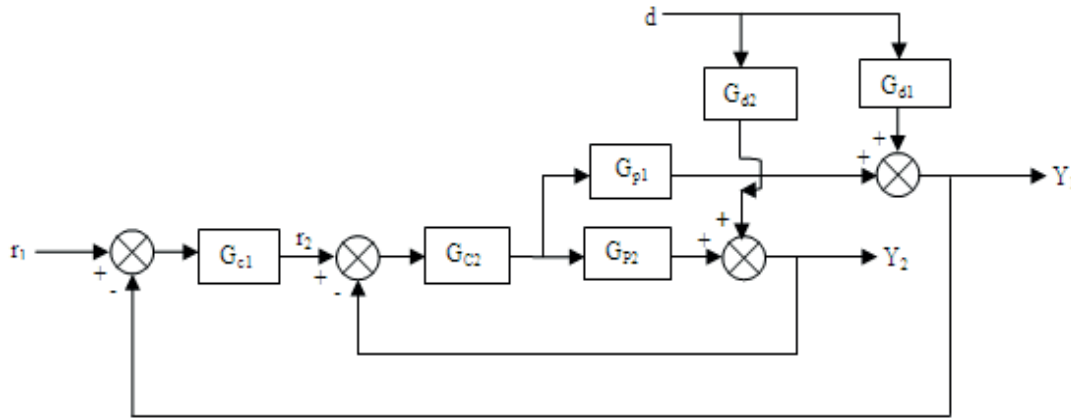


Figure 1. Parallel cascade control system

In parallel cascade control, the manipulated and disturbance variables simultaneously affect primary and secondary outputs. The control objective is to maintain the primary output Y_1 at the set point in the midst of disturbances. Y_2 is controlled by the dynamics of the internal loop. The disturbance rejection in the primary loop depends on the disturbance rejection in the secondary loop as well as its set point tracking. Output of the primary controller, G_{c1} resets the set point of the secondary loop. The secondary measurement, Y_2 is designated to infer the load disturbance. Secondary controller G_{c2} is designed for disturbance rejection purpose, while the primary controller is designed to maintain good servo response. Hence Parallel cascade control structure offers an effective solution to overcome disturbance problems.

The proposed parallel cascade control is shown in Figure 2. In parallel cascade control, the dynamics of the secondary loop is faster than the primary loop. They have negligible or no dead time compared to primary process, which has larger dead time. So secondary controller, G_{c2} is designed based on IMC method. G_{p2m} is the secondary process model and G_{f2} is the set point filter. The primary process are usually non linear with large dead time and they are difficult to model. Non-linear boundary conditions, time varying characteristics and a complex physical environment create many difficulties in designing a

mathematical model. So, the existing PID or IMC based controllers will not be effective. They hardly satisfy a high performance criterion with respect to both parameter insensitivity and disturbance rejection. Also, effectiveness of IMC controller depends on the accuracy of the process model. Though it gives good servo response, its load regulation is poor. On the other hand, fuzzy control is one of the useful control techniques for uncertain and ill-defined nonlinear systems because available qualitative operator and design knowledge can be easily implemented. Hence the existing control can be improved by incorporating fuzzy controller for the primary controller, G_{c1} . Further, the dead time in the outer loop is compensated by incorporating Smith predictor in the primary loop, which compensates the dead time [16]. G_{pm} is the process model of the outer loop without dead time. θ_m is the dead time of the model. By including the filter G_{f1} with time constant τ_f , for the predicted disturbance, the robustness of the delay compensator can be increased [17], where

$$G_{f1} = \frac{1}{\tau_f s + 1} \quad (1)$$

The closed loop transfer functions of secondary and primary loop for the system in Figure 2 are,

for secondary loop (assuming $G_{p2} = G_{p2m}$)

$$Y_2 = G_{p2m} G_{c2} G_{f2} r_2 + (1 - G_{p2m} G_{c2}) G_{d2} d \quad (2)$$



Further, (since $G_{p2} = G_{p2m}$), the relation between primary output (Y_1) and secondary set point (r_2) is obtained as

$$Y_1 = G_{p1} G_{c2} G_{f2} r_2 \quad (3)$$

However, for the primary loop, we also have

$$Y_1 = \frac{G_{f2} G_{p1} G_{c1} G_{c2}}{1 + G_{c1} (G_{pm} (1 - G_{f1} e^{-\theta_m s}) + G_{p1} G_{c2} G_{f1} G_{f2})} r_1 + \frac{(1 + G_{pm} G_{c1} (1 - G_{f1} e^{-\theta_m s})) (G_{d1} - G_{p1} G_{c2} G_{d2})}{1 + G_{c1} (G_{pm} (1 - G_{f1} e^{-\theta_m s}) + G_{p1} G_{c2} G_{f1} G_{f2})} d \quad (4)$$

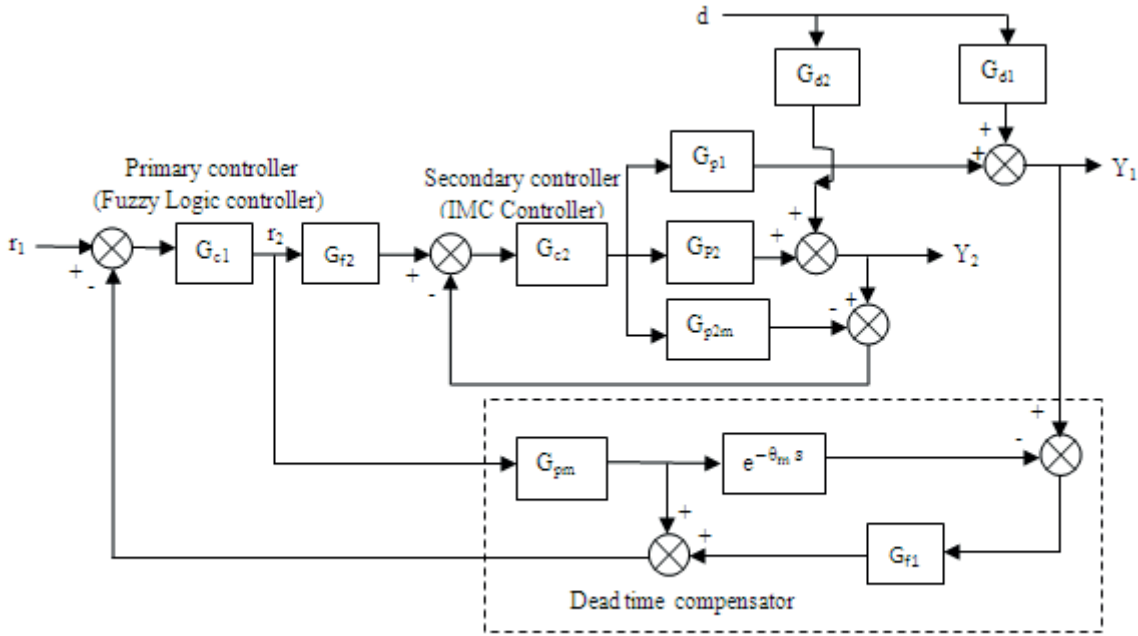


Figure 2. Fuzzy control with Smith predictor implementation for parallel cascade control

From Eq. (3), by assuming perfect model conditions, the primary process model is obtained as

$$G_{pm} = G_{p1} G_{c2} G_{f2} \quad (5)$$

4. Controller Design and Simulation

Simulation results are obtained for different case studies. Experiment is carried out for both slow and fast dynamic systems. The performance of the designed controller is compared with the recently reported methods.

Case study No.1

The following process was modeled by Semino and Bramilla [8]. The process is characterized by the same dynamics in G_{p1} and G_{p2} and in G_{d1} and G_{d2} with exception of a dead time which corresponds to a delay in measurement of the primary output.

The process and disturbance transfer function models are given as

$$G_{p1} = G_{d1} = \frac{1.24e^{-33s}}{30s + 1} \quad (6)$$

$$G_{p2} = G_{d2} = \frac{3.14e^{-9s}}{30s + 1} \quad (7)$$

Here the primary process is dead time dominant and the secondary process is purely lag dominant. Lee et al. [9] have designed the secondary controller for slow dynamics and the primary controller for the fast dynamics and found improved results. For the proposed method the Internal model controller is designed for the secondary controller and fuzzy logic controller is designed for the primary controller.

IMC based Controller Design (G_{c2})

As stated in section I, the secondary controller is designed based on IMC principles [12, 18]. The inner loop transfer function in the form of a First order plus time delay (FOPTD), can be expressed as

$$G_{p2} = \frac{k_2 e^{-\theta_2 s}}{(\tau_2 s + 1)} \quad (8)$$

Where, k_2 is the process gain and τ_2 is the process time constant. Then the secondary controller using IMC principle, is given by



$$G_{c2} = \frac{(\tau_2 s + 1)}{k_2(\lambda s + 1)} \quad (9)$$

where, λ is the only tuning parameter to be found. Smaller the value of λ , the better the performance of the cascade control system. If a faster response is required, λ can be chosen as half time delay of the inner loop, i.e., $\lambda = \theta_2/2$. The lead term in the closed loop response of the secondary output for set point changes may cause an overshoot and large settling time which will affect the closed loop system performance. To eliminate this undesirable overshoot, a set point filter G_{f2} is used [12].

The secondary controller tuning parameter is selected as $\lambda = 5$, so as to achieve the same secondary disturbance rejection as that of Seshagiri Rao et al.[13]. The secondary controller is obtained as

$$G_{c2} = \frac{(30s+1)(14.6s+1)}{3.1(5s+1)^2} \quad (10)$$

and the loop filter

$$G_{f2} = \frac{1}{14.6s + 1} \quad (11)$$

The primary process model is obtained as

$$G_{pm}e^{-\theta_m s} = \frac{0.4e^{-33s}}{(5s + 1)^2} \quad (12)$$

Using Skogestad's half rule [19], the above model is reduced to FOPTD Model and is obtained as

$$G_{pm}e^{-\theta_m s} = \frac{0.4e^{-35.5s}}{(7.5s+1)} \quad (13)$$

Fuzzy Logic Controller Design (G_{cl})

Control algorithms with fuzzy logic controllers (FLC) prove to offer better robustness and efficiency in the case of more complicated non-linear and time varying systems and systems with large dead time in comparison with conventional PID controllers. While a PID controller allows no flexibility of structure, a fuzzy logic controller can be whatever it needs to be [20]. The difficulty in obtaining the optimum settling time of PID controller is mitigated by using fuzzy logic controller which gives the opportunity to describe the control action in quantitative form and symbolic form. Hence, fuzzy logic controller is designed for the primary controller (G_{c1}). It must be recalled that the general design of the fuzzy logic controller proceeds along the following steps:

- (i) Fuzzification
- (ii) Design of Membership function
- (iii) Defuzzification

The design process needs a systematic method for obtaining the rule base and domain ranges. The error (e) and change in error (Δe) are chosen as an input to FLC.

where,

$$\text{Error} = \text{Set point} - \text{Measured Variable} \quad (14)$$

$$\text{Change in Error} = \text{Present Error} - \text{Previous Error} \quad (15)$$

Since the design of FLC is based on "If - Then" Rule-based considerations, it must be noted that in the present context, the two signals given in (12) and (13) constitute the rule-antecedent in the formation of rule base [i.e., the "If"-part]; the control output (u) represents the contents of the rule-consequent in the formation of rule base [i.e., the "Then"-part].

Fuzzification

Fuzzification is the quantization of crisp quantity to fuzzy linguistic variables. In many of the quantities that are considered crisp and deterministic are actually not deterministic at all. Fuzzification is related to the vagueness and imprecision in a natural language. It is a subjective valuation, which transforms a measurement into a valuation of a subjective value. Hence, it could be defined as a mapping from an observed input space to fuzzy sets in certain input universes of discourse. Fuzzification plays an important role in dealing with uncertain information, which might be objective or subjective in nature [21].

Design of Membership function

The ranges of input and output are assigned with linguistic variables. These variables transform the numerical values of the input of the fuzzy controller to fuzzy quantities. These linguistic variables specify the quality of the control. Among all membership functions, triangular membership function gives good results and easy to implement. In this work, the triangular membership function is used for designing the FLC. The designed fuzzy triangles for Error (e), Change in Error (Δe) and the controller output (u) are shown in Figure 3.

The Linguistic variables for error (e) is classified into: Negative big (e_{-big}); Negative Small (e_{-sm}); Zero (e_{zero}); Positive Small (e_{+sm}); Positive Big (e_{+big})

The Linguistic variables for Change in Error (Δe) is classified into:

Negative big (Δe_{-big}); Negative Small (Δe_{-sm}); Zero (Δe_{zero}); Positive Small (Δe_{+sm}); Positive Big (Δe_{+big})

The Linguistic variables for FLC output (u) is classified into:

Negative big (u_{-big}); Negative Small (u_{-sm}); Zero (u_{zero}); Positive Small (u_{+sm}); Positive Big (u_{+big})

Knowledge Base

The knowledge base of an FLC is comprised of two components, a database and fuzzy control base. The concepts associated with a database are used to



characterize fuzzy control rules and a fuzzy data manipulation in FLC. It should be noted that the correct choice of the membership function of a term set plays an essential role in the success of the application. A lookup table based on discrete universes defines the output of a controller for all possible combinations of the input signals. A fuzzy system is characterized by a set of linguistic statements. It is in the form of "IF-THEN" rules. Fuzzy conditional statements form the rules or the rule set of the FLC. The decision table is given in Table 1.

Table 1: Rule Matrix of the proposed FLC

$e(k) \backslash \Delta e(k)$	e_{-big}	e_{-sm}	e_{zero}	e_{+sm}	e_{+big}
Δe_{-big}	U_{-big}	U_{-big}	U_{-big}	U_{-sm}	U_{zero}
Δe_{-sm}	U_{-big}	U_{-big}	U_{-sm}	U_{zero}	U_{+sm}
Δe_{zero}	U_{-big}	U_{-sm}	U_{zero}	U_{+sm}	U_{+big}
Δe_{+sm}	U_{-sm}	U_{zero}	U_{+sm}	U_{+big}	U_{+big}
Δe_{+big}	U_{zero}	U_{+sm}	U_{+big}	U_{+big}	U_{+big}

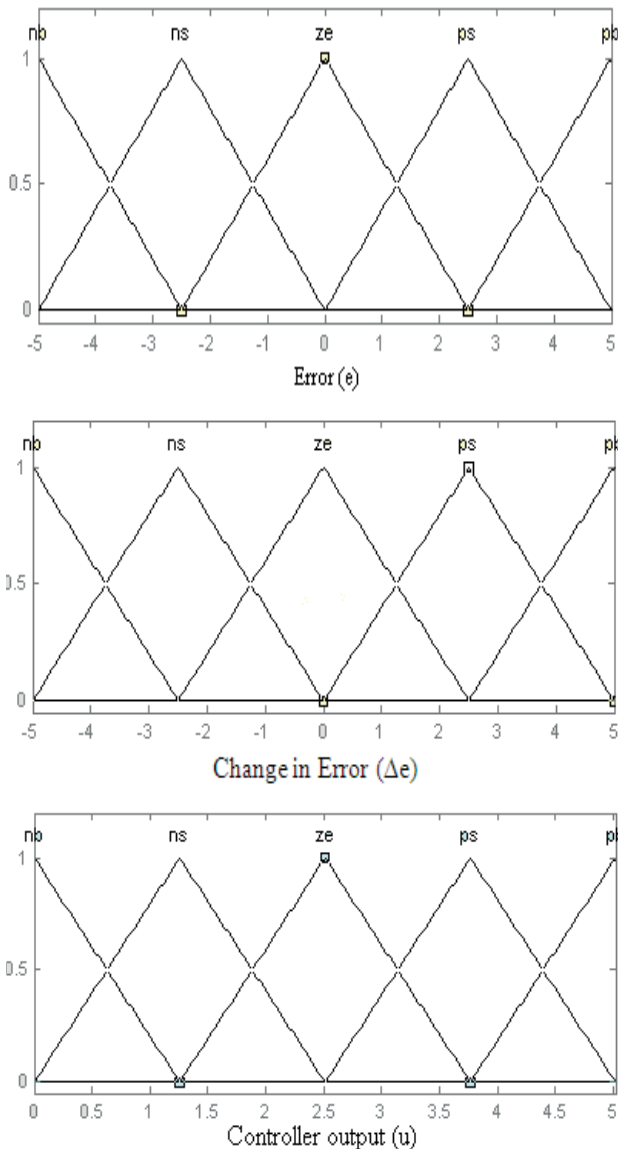


Figure 3: Fuzzy triangles for Error (e), Change in Error (Δe) and Controller output (u) for Case study No. 1.

Defuzzification

The final stage of fuzzy logic implementation is defuzzification. In this process, the fuzzy values are converted into acceptable crisp values. Centroid defuzzification is used in the majority of the cases to develop the expected value for each of these output variables [22].

By the centroid defuzzification method, the crisp controller output (u) is obtained by,

$$u = \frac{\sum_{i=1}^N f_i \mu_i}{\sum_{i=1}^N \mu_i} \quad (16)$$

where,

u = Crisp valued controller output

f_i = Fuzzy valued controller output

μ_i = Input membership function

N = Total number of rules

From the above formulae, controller output is obtained which is given to the secondary process. By this way, fuzzy logic controller is designed and implemented for the primary controller.

With the above controller design, the proposed method is simulated by giving unit step change in the set point at time, $t=0s$ and a negative step input of magnitude 2 in the load disturbance at $t=350s$ respectively. The closed loop response is shown in Figure 4. The closed loop response of the proposed method reaches the set point with less rise time and settles to the final steady state value with minimum overshoot when compared to that of [13]. The values obtained are listed in Table 2. Also, it is observed that the disturbance rejection of the proposed controller is better in terms of fast recovery and minimum overshoot.



To analyze the robustness of the controller, a perturbation of +20% in the primary process time delay and +20% of disturbance time delay are given to the model. The corresponding closed loop response is shown in Figure 5. For obtaining the result analysis under transient conditions rise time (t_r), peak overshoot (%OS) and settling time (t_s) are compared. For obtaining the result analysis under steady state conditions integral square error (ISE) and integral of absolute error (IAE) are compared.

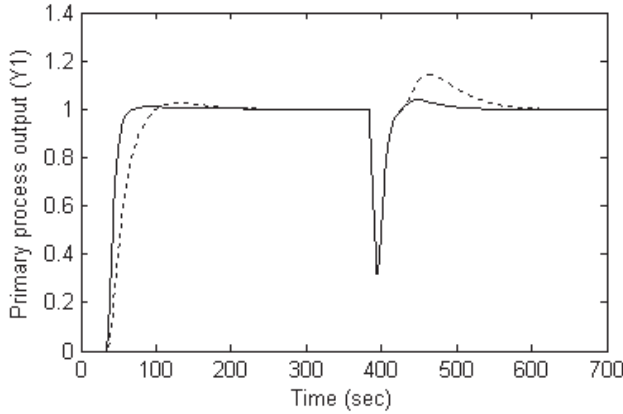


Figure 4: Closed loop response for Case study No.1 for perfect model. Solid – Proposed method, dash – Seshagiri Rao et al. [13].

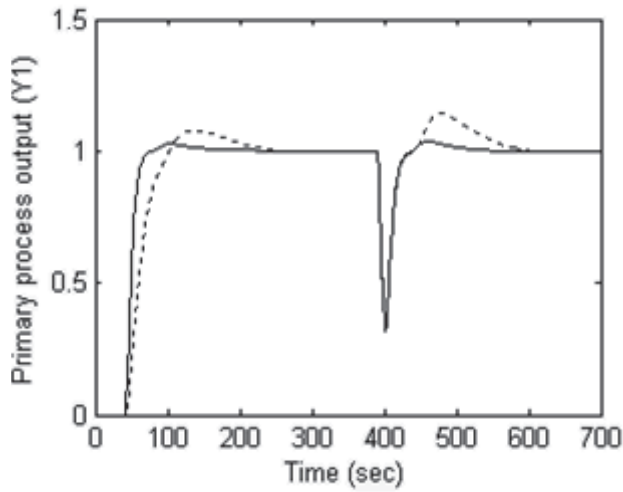


Figure 5: Closed loop response for Case study No.1 for perturbation of 20% in the primary process and disturbance delay. Solid – Proposed method, dash – Seshagiri Rao et al. [13].

Table 2: Performance analysis for Case study No.1

	t_r (s)	%OS	t_s (s)	ISE	IAE
Proposed Method	15.8	1	60	3.78	17.28
Seshagiri Rao et al	31.9	2.6	90	12.64	26

The values of these different performance indices are provided in Table 2. There is 51% improvement in rise time, 33 % improvement in settling time and

the peak overshoot is reduced by 62% when compared to [13]. The ISE and IAE of the proposed method are also reduced considerably when compared with existing methods. Similar improvements were obtained for disturbance rejection also. Seshagiri Rao et al. [13] have already shown that their method performs better than Lee et al. [12].

Case study No.2

Here, an example from Luyben [3] is considered. The dead time in the primary process and disturbance is considered as 4 times the time constant. The resulting process and disturbance transfer function models are given as

$$G_{p1} = G_{d1} = \frac{e^{-80s}}{20s+1} \quad (17)$$

$$G_{p2} = G_{d2} = \frac{1}{10s+1} \quad (18)$$

Here the secondary process dynamics is faster than the desired closed loop response. The tuning parameter for the secondary controller is selected as $\lambda = 1$, so as to achieve the same secondary disturbance rejection as that of [13].

The secondary controller is obtained as

$$G_{c2} = \frac{(10s+1)(1.9s+1)}{(s+1)^2} \quad (19)$$

And the loop filter

$$G_{f2} = \frac{1}{1.9s+1} \quad (20)$$

The primary process model is obtained as

$$G_{pm}e^{-\theta_m s} = \frac{(10s+1)e^{-80s}}{(20s+1)(5s+1)^2} \quad (21)$$

Using Skogestad's half rule, the above model is reduced to FOPTD Model and is obtained as

$$G_{pm}e^{-\theta_m s} = \frac{e^{-80.5s}}{(20.5s+1)} \quad (22)$$

The primary fuzzy controller is designed as explained above. The designed fuzzy triangles for Error (e), Change in Error (Δe) and the controller output (u) are shown in Figure 6. With the above controller design, the proposed method is simulated by giving unit step change in the set point at time, $t=0s$ and a negative step input of magnitude 10 in the load disturbance at $t=250s$ respectively. The closed loop response is shown in Figure 7.



Here too, the closed loop response of the proposed method reaches the set point with less rise time and settles to the final steady state value with minimum overshoot when compared to that of [13]. Also, disturbance rejection of the proposed controller is better and the system recovers faster and reaches the set point with minimum overshoot. Hence it is observed that the proposed controller gives better response and the control action of the proposed method was found to be superior to that of [13]. The values of different performance indices are provided in Table 3. There was 44% improvement in rise time and 3% improvement in settling time. There is no peak overshoot when compared to that of [13]. The ISE and IAE of the proposed method are also very minimum and comparable with existing methods. Similar improvements were obtained for disturbance rejection also.

Table 3: Performance analysis for Case study No.2

	t_r (s)	%OS	t_s (s)	ISE	IAE
Proposed Method	16.5	0	114	6.89	19.53
Seshagiri Rao et. al	29.4	8.5	118	14.69	28.91

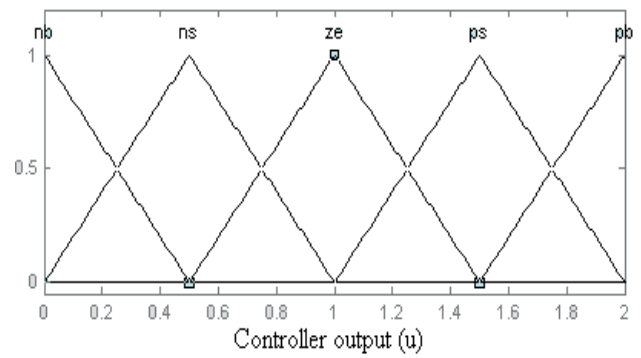
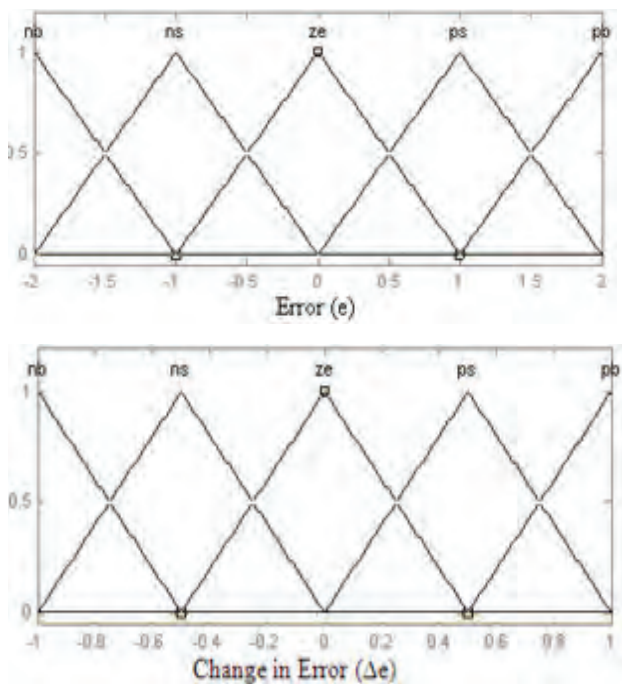
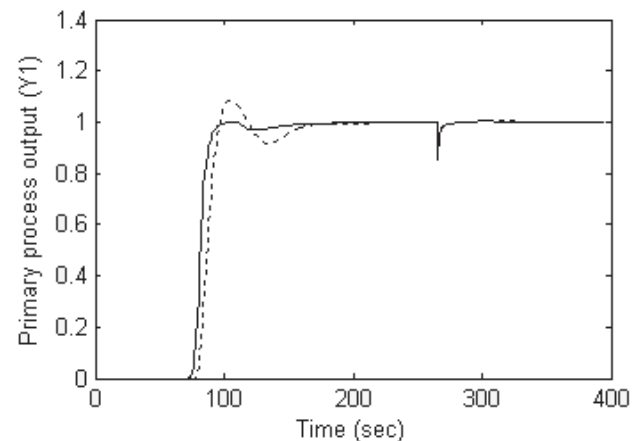
Figure 6: Fuzzy triangles for Error (e), Change in Error (Δe) and Controller output (u) for Case study No. 2

Figure 7: Closed loop response for Case study No.2 for perfect model. Solid – Proposed method, dash – Seshagiri Rao et al. [13].

5. Conclusion

Fuzzy controller is designed and implemented for the primary controller of parallel cascade control system. Dead time compensator is incorporated in the primary loop. The proposed structure brings together the best merits of the fuzzy control and the parallel cascade control structure. The performance of the proposed method is analyzed for both slow and fast dynamic systems and it is shown by the case studies, that the fuzzy based parallel cascade control gives a better performance for both set point tracking and load disturbance rejection. The result is a stable control system with fast reaction time and the proposed method outperforms the recently reported methods in performance measures for both transient and steady state conditions.



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Geometrical Non-Linear Steady State Forced Periodic Vibration of Fully Clamped Composite Laminated Rectangular Plates: a Multi Mode Approach

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Abstract

A model for the steady state, geometrically non-linear, periodic forced vibration of fully clamped thin rectangular composite plates under harmonic external excitation is presented. The equations of motion have been derived using spectral analysis, Lagrange's equations, and the harmonic balance method (HBM). This forced case has been examined using a single and multimode approach. A multidimensional Duffing equation in the case of fully clamped rectangular laminated composite plates has been obtained. The behaviour of the symmetrically laminated plates on the responses to forced vibration under harmonic distributed force. Results obtained are compared with some solutions based on hierarchical finite element method. It is found that accurate results may be produced using the present model.

Keywords: *Clamped thin rectangular composite plates, External excitation, Harmonic balance method, Multimode approach*

1. Introduction

With the increasing use of laminated composite materials in structures, engineers and researchers are faced with challenging problems associated with the understanding of the behaviour of these new materials. The correct and effective use of them requires more complex analysis tools in order to predict accurately their response to external loadings. As most modern laminated plates are

made of high performance materials, they are usually used in severe working conditions resulting in large vibration amplitudes response. This means that the effect of geometric non-linearity needs to be included. A considerable amount of research work has been carried out on geometrically non-linear behaviour of laminated composite plates, particularly free vibration of thin plates [1 to 10]. However, attention has been also devoted to the forced vibration of composite plates, as may be seen by the amount of published works on the subject. In reference [11 to 13] the finite element method has been extended to investigate large amplitude forced vibrations of rectangular composite plates. Based on the hierarchical finite element method (HFEM), a forced vibration analysis of symmetrically laminated plates has been studied in reference [14]. The loads considered are harmonic acoustic plane waves impinging on the plate surface in normal direction and at grazing incidence. Only linear vibrations have been analysed. Recently, a model for the steady state, geometrically non-linear, periodic forced vibration of both isotropic and composite laminated rectangular plates under harmonic external excitation has been presented in [15]. The model has been derived by applying the principle of virtual work and the HFEM. The equations obtained have been transformed into the frequency domain by the HBM and have been solved by a continuation method. In the above reference, the convergence properties of the model have been discussed by applying it to isotropic and



to composite laminated plates. The stability study and analysis of multi-modal response of the model have been investigated in reference [16].

In a previous work [17] a model for the steady state, geometrically non-linear, periodic forced vibration of fully clamped thin rectangular composite plates, under harmonic external excitation was presented. The model has been derived using spectral analysis, Lagrange's equations, and the harmonic balance method (HBM). A multidimensional Duffing equation in the case of fully clamped rectangular laminated composite plates has been obtained. It was assumed for simplicity only one mode, this set reduces to the Duffing equation, very well known in one mode analyses of non-linear systems having cubic non-linearities. The frequency response curves have been obtained at the plate centre, for various levels of loading. It appeared that the method works well, since excellent agreement is found between the result of the present model, and those published in the literature in the centre of the plate. In this paper, the model developed in [17] is extended to investigate the multimode response. The response of the plates to external harmonic excitations is analysed and compared with some published results. Fully clamped boundaries are considered because they are adequate to model many real panel-type situations, such as aircraft wing panels [18]. Results are given for single and multimode approach.

In this paper we will rely on the Lagrange equations to study the vibrations of a rectangular plate and built a composite material of known characteristics. Then we will use Harmonic Balance Method coupled with the routine NS01A to develop the method mentioned above. And finally we will compare results with other studies that already exist

2. Theory

Consider a thin plate of thickness H having a length a and a width b , and composed of thin orthotropic layers, layered along length and width, that is x and y axis, alternately. The origin of the Cartesian co-ordinate system is located in one corner of the mid-plane with the z axis being perpendicular to this plane as shown in Figure 1. Assuming that the plate undergoes large amplitudes, the Von-Kármán type strain-displacement relationships as given below are used:

$$\{\varepsilon\} = \{\varepsilon^0\} + z\{\kappa\} + \{\lambda^0\} \quad (1)$$

in which $\{\varepsilon^0\}$, $\{\kappa\}$ and $\{\lambda^0\}$ are given by:

$$\{\varepsilon^0\} = \begin{bmatrix} \varepsilon_x^0 \\ \varepsilon_y^0 \\ \gamma_{xy}^0 \end{bmatrix} = \begin{bmatrix} \frac{\partial U}{\partial x} \\ \frac{\partial V}{\partial y} \\ \frac{\partial U}{\partial y} + \frac{\partial V}{\partial x} \end{bmatrix} \quad (2)$$

$$\{\kappa\} = \begin{bmatrix} \kappa_x \\ \kappa_y \\ \kappa_{xy} \end{bmatrix} = \begin{bmatrix} -\frac{\partial^2 W}{\partial x^2} \\ -\frac{\partial^2 W}{\partial y^2} \\ -2\frac{\partial^2 W}{\partial xy} \end{bmatrix} \quad (3)$$

$$\{\lambda^0\} = \begin{bmatrix} \lambda_x^0 \\ \lambda_y^0 \\ \lambda_{xy}^0 \end{bmatrix} = \begin{bmatrix} \frac{1}{2} \left(\frac{\partial W}{\partial x} \right)^2 \\ \frac{1}{2} \left(\frac{\partial W}{\partial y} \right)^2 \\ \frac{\partial W}{\partial x} \cdot \frac{\partial W}{\partial y} \end{bmatrix} \quad (4)$$

U , V and W are the displacements of the plate mid-plane, in the x , y and z directions respectively.

For the laminated plate having n layers shown in Figure 2, the stress in the k_{th} layer can be expressed in terms of the laminate middle surface strains and curvatures as:

$$\{\sigma_k\} = [\bar{Q}]_k \cdot \{\varepsilon\} \quad (5)$$

In which $\{\sigma\}_k^T = [\sigma_x \quad \sigma_y \quad \tau_{xy}]$ and terms of matrix $[\bar{Q}]$ can be obtained by the relationships given in [19].

The force and moment resultants for the k_{th} layer are defined as follows:

$$(N_x^{(k)}, N_y^{(k)}, N_{xy}^{(k)}) = \int_{h_{k-1}}^{h_k} (\sigma_x^{(k)}, \sigma_y^{(k)}, \sigma_{xy}^{(k)}) dz \quad (6)$$

$$(M_x^{(k)}, M_y^{(k)}, M_{xy}^{(k)}) = \int_{h_{k-1}}^{h_k} (\sigma_x^{(k)}, \sigma_y^{(k)}, \sigma_{xy}^{(k)}) z dz \quad (7)$$

Where h_k is the distance from the mid-plane to the surface of the k_{th} layer.

The in-plane forces and bending moments in a plate are given by:

$$\begin{bmatrix} N \\ M \end{bmatrix} = \begin{bmatrix} A & B \\ B & D \end{bmatrix} \begin{bmatrix} \{\varepsilon^0\} + \{\lambda^0\} \\ \{\kappa\} \end{bmatrix} \quad (8)$$



A, B and D are the symmetric matrices given by (9).
[B] = 0 for symmetrically laminated plates [20].

$$(A_{ij}, B_{ij}, D_{ij}) = \int_{-\frac{H}{2}}^{\frac{H}{2}} Q_{ij}^{(k)}(1, z, z^2) dz \quad (9)$$

Where the $Q_{ij}^{(k)}$ are the reduced stiffness coefficients of the k_{th} layer in the plate co-ordinates. The expression for the bending strain energy V_b , axial strain energy V_a and kinetic energy T are given by [21]:

$$V_b = \frac{1}{2} \int_S \left\{ D_{11} \left(\frac{\partial^2 W}{\partial x^2} \right)^2 + 2D_{12} \frac{\partial^2 W}{\partial y^2} \frac{\partial^2 W}{\partial x^2} + D_{22} \left(\frac{\partial^2 W}{\partial y^2} \right)^2 + 4D_{16} \frac{\partial^2 W}{\partial x^2} \frac{\partial^2 W}{\partial xy} + 4D_{26} \frac{\partial^2 W}{\partial y^2} \frac{\partial^2 W}{\partial xy} + 4D_{66} \left(\frac{\partial^2 W}{\partial xy} \right)^2 \right\} dS \quad (10)$$

$$V_a = \frac{1}{2} \int_S \left\{ \frac{A_{11}}{4} \left(\frac{\partial W}{\partial x} \right)^4 + \frac{A_{22}}{4} \left(\frac{\partial W}{\partial y} \right)^4 + \left[\frac{A_{12}}{2} + A_{66} \right] \left(\frac{\partial W}{\partial y} \right)^2 \left(\frac{\partial W}{\partial x} \right)^2 + A_{16} \left(\frac{\partial W}{\partial x} \right)^3 \frac{\partial W}{\partial y} + A_{26} \left(\frac{\partial W}{\partial y} \right)^3 \frac{\partial W}{\partial x} \right\} dS \quad (11)$$

and

$$T = \frac{1}{2} \rho H \int_S \left(\frac{\partial W}{\partial t} \right)^2 dS \quad (12)$$

Where S is the plate surface $[0,a] \times [0,b]$ and dS is the elementary surface $dx dy$.

In the above expressions, the assumption of neglecting the inplane displacements U and V in the energy expressions has been made as for the fully clamped rectangular isotropic plates analysis considered in [22,23]. The range of validity of this assumption has been extensively discussed in the light of the experimental and numerical results obtained for the frequency amplitude dependence and the bending stress estimates obtained at large vibrations amplitudes. The results obtained via this assumption were compared with the previous ones based on various methods such as the finite element method, the method based on Berger's approximation, the ultraspherical polynomial method and the elliptic function method. It was found that the percentage error in the non-linear frequency estimates based on this assumption, for amplitudes up to 1.5 times thickness, does not

exceed 1.3 %. Also, in the experimental investigation of the non-linear behaviour of fully clamped rectangular plates at large vibration amplitudes presented in reference [24], it was found that the rate of increase in bending stresses estimates, obtained from measured data, was in very good agreement with that obtained from the theory, in which the assumption of zero in-plane displacements was made. The above references concerned with the isotropic case. It has led to the conclusion that this assumption allows a great simplification in the modelling and a great reduction in the computation time when calculating the non-linear mode shapes, the associated non-linear frequencies, and the non-linear bending stress patterns, for a reasonable range of vibrations amplitudes and plates aspect ratios. In the present work, concerned with symmetrically laminated plates, the validity of this assumption has also been examined, via comparison of the non-linear frequency estimates it leads to, with those obtained by the HFEM, for two symmetrically laminated plates, with different lay-up and thicknesses.

If the time and space functions are supposed to be separated and harmonic motion is assumed, and the generalised parameterisation and usual summation convention defined in [25], the transverse displacement can be written as:

$$W(x, y, t) = q_i(t) w_i(x, y) \quad (13)$$

where $w_i(x, y)$ are the basic functions. Substituting W in the expressions (10-12) for V_a , V_b and T , the discretisation of the kinetic energy and the total strain energy V leads to:

$$V_b = \frac{1}{2} q_i q_j k_{ij} \quad (14)$$

$$V_a = \frac{1}{2} q_i q_j q_k q_l b_{ijkl} \quad (15)$$

$$T = \frac{1}{2} q_i q_j \omega^2 m_{ij} \quad (16)$$

where the terms m_{ij} , k_{ij} and b_{ijkl} are given as in reference [6 to 9,25], and indices i, j, k and l are summed over 1 to n

2.1 Non-linear Free vibration

The numerical model developed in [22,26,27], and used in [6 to 10], by applying Hamilton's principle to the free response problem, can also be derived



using Lagrange's equations and the Harmonic Balance Method (HBM).

For a conservative system, having n degrees of freedom corresponding to n parameters $q_r(t)$, $r = 1$ to n , Lagrange's equations can be written, if no forcing term is considered, as [28]:

$$-\frac{\partial}{\partial t} \frac{\partial T}{\partial \dot{q}_r} + \frac{\partial T}{\partial q_r} - \frac{\partial V}{\partial q_r} = 0 \quad r = 1 \text{ to } n \quad (17)$$

Substituting equations (14-16) in (17) leads to the following set of non-linear differential equations, similar to that found in [6 to 10,22,29,30], and considered as a multidimensional form of the very well known Duffing equation:

$$\ddot{q}_i m_{ir} + q_i k_{ir} + 2q_i q_j q_k b_{ijk} = 0 \quad r=1 \text{ to } n \quad (18)$$

which can be written in matrix form as:

$$[M]\{\ddot{q}\} + [K]\{q\} + 2[B(\{q\})]\{q\} = \{0\} \quad (19)$$

where $[M]$, $[K]$, $[B]$ and $\{q\}$ are the mass, the linear rigidity, the non-linear rigidity tensors and the column vector of generalised parameters $\{q\}^T = [q_1, q_2, \dots, q_n]$ respectively.

By assuming that harmonic motion takes place:

$$q_i(t) = a_i \cos(\omega t) \quad (20)$$

Inserting equation (20) into equation (19), and applying the HBM one arrives at the following system of equations:

$$2([K] - \omega^2[M])\{A\} + 3[B(\{A\})]\{A\} = \{0\} \quad (21)$$

in which $\{A\}$ is the column vector of the basic function contribution coefficients $\{A\}^T = [a_1, a_2, \dots, a_n]$.

In order to formulate the vibration problem of fully clamped-clamped rectangular laminated plates corresponding to large deflections in non-dimensional terms, we put, as in reference [6,25]:

$$w_i(x, y) = H w_i^* \left(\frac{x}{a}, \frac{y}{b} \right) = H w_i^*(x^*, y^*) \quad (22)$$

in which $w_i(x, y)$ is the i th mode of the plate considered.

The non-dimensional axial and bending stiffnesses A_{ij}^* and D_{ij}^* :

$$A_{ij}^* = \frac{A_{ij}}{HE} \quad D_{ij}^* = \frac{D_{ij}}{H^3 E} \quad (23, 24)$$

The non-dimensional tensors k_{ij}^* , b_{ijkl}^* and m_{ij}^* can be obtained as:

$$k_{ij} = \frac{aH^5 E}{b^3} k_{ij}^* \quad b_{ijkl} = \frac{aH^5 E}{b^3} b_{ijkl}^* \quad m_{ij} = \rho H^3 a b m_{ij}^* \quad (25-27)$$

and:

$$\omega^2 = \frac{H^2 E}{\rho b^4} \omega^{*2} \quad (28)$$

Inserting equations (22-28) into equation (21), the following equations is obtained:

$$([K^*] - \omega^{*2}[M^*])\{A\} + \frac{3}{2}[B^*(\{A\})]\{A\} = \{0\} \quad (29)$$

which may be written also, using tensor notation as:

$$-\omega^{*2} a_i m_{ir}^* + a_i k_{ir}^* + \frac{3}{2} a_i a_j a_k b_{ijk}^* = 0 \quad r = 1 \text{ to } n \quad (30)$$

For the non-linear problem (29), as the modulus of $\{A\}$ is also necessary to determine the physical amplitude of vibration, we have $(n+1)$ unknowns which are the n coefficients of $\{A\}$ and ω . A further equation has to be added to equations (29) in order to complete the formulation. As no dissipation is considered here, such an equation can be obtained by applying the principle of conservation of energy, which can be written, as in reference [25]:

$$V_{\max} = T_{\max} \quad (31)$$

This leads to the following equation:

$$\omega^{*2} = \frac{a_i a_j k_{ij}^* + a_i a_j a_k a_l b_{ijkl}^*}{a_i a_j m_{ij}^*} \quad (32)$$

To obtain the first non-linear mode shape of the plate considered, the contribution of the first basic function is first fixed and the other basic function contributions are calculated via numerical solution



of the remaining (n-1) non-linear algebraic equations.

2.2 Non-linear forced vibration

The model developed above is extended in this subsection to the steady-state periodic forced response case. The excitations considered will be of the form $F(x,y,t) = F \cos \omega t$. Numerical, experimental and analytical investigations confirm that plates vibrating with displacement amplitudes of the order of their thickness due to harmonic excitation forces, usually experience periodic motion [29]. In this work, only the first harmonic motions is examined, via use of the HBM.

Consider a plate subjected to a transverse concentrated force F_c applied at the point (x_0, y_0) , or to a transverse distributed harmonic uniform force F_d , distributed over the range S , where S is the surface of the plate or a part of it, F_c and F_d are given by:

$$F^c(x, y, t) = F^c \cos(\omega t) \delta(x - x_0) \delta(y - y_0) \quad (33)$$

$$F^d(x, y, t) = F^d \cos(\omega t) \quad (34)$$

δ is the Dirac function. The corresponding generalised forces $F_i^c(t)$ and $F_i^d(t)$ can be written as:

$$F_i^c(t) = F^c \cos(\omega t) w_i(x_0, y_0) = f_i^c \cos(\omega t) \quad (35)$$

$$F_i^d(t) = F^d \cos(\omega t) \int_S w_i(x, y) dx dy = f_i^d \cos(\omega t) \quad (36)$$

The non-dimensional generalised forces f_i^{c*} and f_i^{d*} corresponding to a concentrated force at (x_0, y_0) , and a uniformly distributed force are given by:

$$f_i^{c*} = F^c \frac{b^3}{aEH^4} w_i^*(x_0, y_0) \quad (37)$$

$$f_i^{d*} = F^d \frac{b^4}{EH^4} \iint_{S^*} w_i^*(x^*, y^*) dx^* dy^* \quad (38)$$

Adding now the forcing term $\{F(t)\}$ to the right hand side of equation (19) leads to:

$$[M]\{\ddot{q}\} + K\{q\} + 2[B(\{q\})]\{q\} = \{F(t)\} \quad (39)$$

Inserting equation (20) into (39) and applying the HBM one obtain the following system of equations:

$$2([K] - \omega^2[M])\{A\} + 3[B(\{A\})]\{A\} = \{F(t)\} \quad (40)$$

Inserting equations (22-28) into equation (40) leads to:

$$([K^*] - \omega^{*2}[M^*])\{A\} + \frac{3}{2}[B^*(\{A\})]\{A\} = \{f^*(t)\} \quad (41)$$

Equation (41) can be written using the tensor notation as:

$$-\omega^{*2} a_i m_{ir}^* + a_i k_{ir}^* + \frac{3}{2} a_i a_j a_k b_{ijk}^* = f_i^* \quad (42)$$

This system of equations is similar to that obtained in equation (30) in the free case, in which i varied from 2 to n because the first contribution a_1 was assigned. In equation (42), all of the n equations have a right hand side representing the forcing term f_i^* .

In equation (42), if only one mode is assumed, and k_{ir}^* for $i \neq r$ is assumed to be negligible compared with k_{rr}^* for simplicity one obtains:

$$(k_{11}^* - \omega^{*2} m_{11}^*) a_1 + \frac{3}{2} b_{1111}^* a_1^3 = f_1^* \quad (43)$$

in which m_{11}^* , k_{11}^* and b_{1111}^* are the mass rigidity, and non-linear rigidity terms corresponding to the first function respectively.

Putting

$$\omega_L^{*2} = \frac{k_{11}^*}{m_{11}^*} \quad (44)$$

Equation (43) can be written as:

$$\left(\frac{\omega^*}{\omega_L^*}\right)^2 = 1 + \frac{3}{2} \frac{b_{1111}^*}{k_{11}^*} a_1^2 - \frac{1}{k_{11}^*} \frac{f_1^*}{a_1} \quad (45)$$

3. One dimension non-linear frequency response functions

3.1 Plate studied

Composite laminated plates are widely used in military aircrafts, particularly in smaller size planes [30].

The composite laminated plate (Graphite/epoxy) studied in reference [15] has been used in this work in order to compare the results obtained by the present model with those published in the same reference. The plate has five layers, with the



following orientation of principal axes (45°, -45°, 45°, -45°, 45°). Its dimensions are $a = 300$ mm, $b = 300$ mm, $h = 1$ mm, and each layer has the material properties $E_L = 173$ GN/m², $E_T = E_L/15.4$ GN/m², $G_{LT} = 0.79 E_T$ GN/m², $\nu_{LT} = 0.3$, $\rho = 1540$ kg/m³.

3.2 Numerical results

The single mode assumption reduces the multi-degree-of-freedom system to a single degree of freedom one. It has been shown in previous studies that such an assumption may not be very accurate due to the increase of curvature near the clamps in the case of clamped-clamped beams [26]. But, as will be shown later, the single mode approach remains a good tool for the estimation of the non-linear frequency versus the amplitude of vibration, in regions far from the clamps, and close to the centre. By assuming only one mode, the forced Duffing equation is obtained; the simplicity of this equation and the good estimation of the parameters of interest, lead one to use this approach in spite of the limitation mentioned above. Computing the coefficients of equation (45) for the composite plate for the same dimensions leads to the analytical formulation as:

$$\left(\frac{\omega^*}{\omega_1^*}\right)^2 = 1 + 3.267279a_1^2 + 0.00694146\frac{f_1^*}{a_1} \quad (46)$$

In order to compare our results with those published in reference [15], we put $a_1 = (W_1/h)/w^*$

where w_i^* are the fully clamped rectangular plate functions defined as

$$w_i^*(x^*, y^*) = \frac{1}{G} f_{\alpha i}^{a^*}(x^*) f_{\beta i}^{a^*}(y^*)$$

normalisation scaling factor given by

$$G = \sqrt{\int_{S^*} (f_{\alpha i}^{a^*}(x^*) f_{\beta i}^{a^*}(y^*))^2 dx^* dy^*}$$

and the chosen basic functions $f_i^{a^*}$ are the linear clamped-clamped beam functions, in which v_i , for $i = 1, 2, \dots, n$, are the eigenvalue parameters for a clamped-clamped beam.

$$f_i^{a^*}(x) = \frac{ch(v_i x/L) - \cos(v_i x/L)}{chv_i - \cosv_i} - \frac{sh(v_i x/L) - \sin(v_i x/L)}{shv_i - \sinv_i}$$

The values of the parameters v_i may be computed by solving numerically the transcendental equation $\cosh v_i \cos v_i = 1$. W_1/h is the amplitude ratio at the centre of the plate for the fundamental mode,

and a_1 is the contribution coefficient to the plate deflection of the first plate function.

Equation (46) corresponding to the 1-D non-linear frequency response, for fully clamped laminated plates is represented in Figures 3-4. Figure 3 represents a comparison between results obtained by the present model with those obtained using the hierarchical finite element method [15]. The fully clamped laminated composite plate is excited by a distributed force with a value of 4 N/m² amplitude. Under these conditions, the single mode approach gives a reasonable approximation for the amplitude of vibration of the plate at point $(x^*, y^*) = (0.5, 0.5)$. In Figure 4, the comparison gives a bad approximation for the amplitude of vibration of the plate at point $(x^*, y^*) = (0.75, 0.75)$ when the amplitude ratio exceeds the value of 0.1. To remedy to this situation, and in order to increase the accuracy of the frequency-amplitude relationships, the multimode method used for the free vibration in [6] has been extended to the forced case. The method consists of taking into account all the coordinates instead of only the single resonant coordinate and the equation obtained is a multidimensional Duffing equation. It can be noted here that the model used in reference [15] includes the in-plane, out of-plane functions and the terms containing the three first harmonics.

4. Multi mode approach: Numerical details and results

4.1. Numerical details

The problem consists of solving the non-linear algebraic system (42). This system has been solved numerically using the Harwell library routine NS01A. This routine is based on a hybrid iteration method combining the step descent and Newton's method, which do not require a very good initial estimate of the solution [31]. The step procedure adopted consists on varying the frequency parameter, which allowed solutions to be obtained with quite a small number of iterations (an average of 104 iterations for 18 equations). The fully clamped laminated composite plate is excited by a harmonic distributed force with an amplitude of 4 N/m². The results were calculated using eighteen plate functions.

4.2 Numerical results and discussion

In order to estimate the accuracy of the results obtained by the present theory, and the effects of the approximations adopted, a comparison has been made with previous results, for large forced vibration amplitudes, based on the HFEM presented in [16].



The ω_n^*/ω_l^* calculated from the present model appear to be very close to the results produced by the HFEM which used the in-plane and the three harmonics in the model. Figures 5-9 show a reasonable agreement between results given in reference [16] and the present results for the fully clamped CFRP rectangular plate. In figure 5, the middle-plane in plane displacements are not considered ($p_i = 0$) in reference [16]; the values of the frequency response functions produced by the present model and that obtained by using HFEM are very similar. The comparison with the same values obtained by the model with that published in reference [16] considering the in-plane displacements and plotted in figure 6 show a slightly difference. Under the same conditions ($p_i = 10$, $p_o = 5$), three harmonics have been used in the model described in reference [16], the comparison with the results obtained by the present model in figure 7 show no difference compared to figure 6. This has been explained in reference [16] that the one harmonic approximation gives a reasonable approximation for the amplitude of vibration of the plate at point $(x^*, y^*) = (0, 0)$. However the interest has been focused near the clamps because the non-linearity is very important. Figure 8 shows the comparison between the transverse displacement versus ω_n^*/ω_l^* obtained by the present model and that published in ref. [16] of the fully clamped rectangular laminated plate at the point $(x^*, y^*) = (0.75, 0.75)$. It can be noticed here that there is no difference between the values. For the same plate and conditions, the in-plane displacements have been introduced ($p_i = 10$) in the model of HFEM [16], the resonance frequency increases more with the first harmonic amplitude compared with figure 8. Now, when the three harmonics have been taken into account and introduced in the model of HFEM [16], the results obtained have been compared with that calculated using the present model and plotted in figure 9. It can be seen very clearly that good agreement has been found.

This good agreement shows that the assumption for neglecting U and V in the present model can lead to reasonable estimation of the forced dynamic behaviour of composite plates.

5. Conclusion

A model for geometrical non-linear, steady state, periodic forced multimode vibration of composite laminated plates has been developed by using the Lagrange's equations, and the harmonic balance method. In this work, a single mode approach has

been adopted. The results obtained indicate that this assumption can lead to good results in the centre of the plate. The accuracy of the model decreases near the clamps. To remedy to this situation, the multimode approach method has been used for increasing the accuracy of the frequency-amplitude relationships. The method consists of taking into account all the co-ordinates instead of only the single resonant co-ordinate. The equation obtained is a multidimensional Duffing equation, similar to that obtained for beams in [31], and for isotropic plates in [32]. The non-linear algebraic system has been resolved numerically and the results have been compared to that obtained by using HFEM. The comparison has been shown a good agreement especially near the clamps. Another point that should be made is that the model using HFEM in reference [16] take into account the in-plane displacements and the three harmonics; and the present model the in-plane displacements have been neglected and only one harmonic has been taken into account, these assumptions have extensively discussed in earlier publications [6 to 10, 17, 22 to 27]. Therefore, the good agreement between the results of the present model and that published indicates that the present model not only can successfully be applied to geometrically non-linear free analyses of symmetrically laminated plates, but can also produce accurate estimation and results of the non-linear forced behaviour of laminated plates. A further investigation considering the in-plane displacements and the three harmonics in the model is needed for more accurate results and a better understanding of geometrically non-linear dynamic behaviour.

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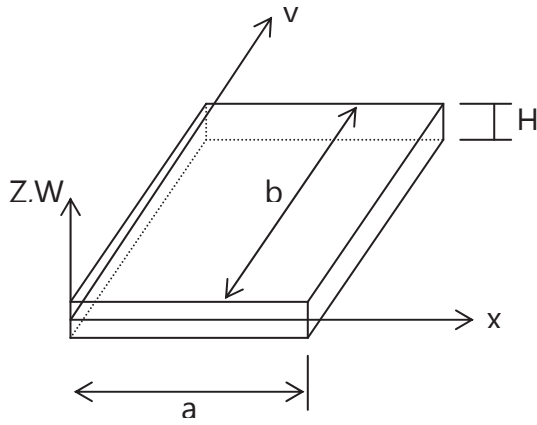


Figure 1: Plate notation

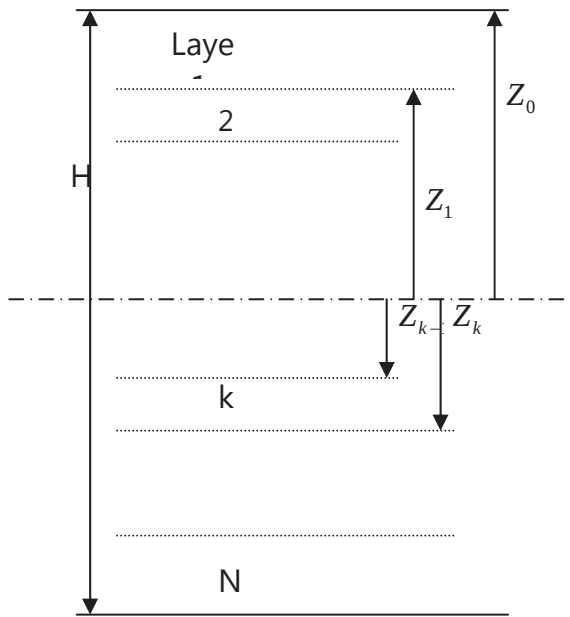


Figure 2: Geometry of n layered laminated plate.

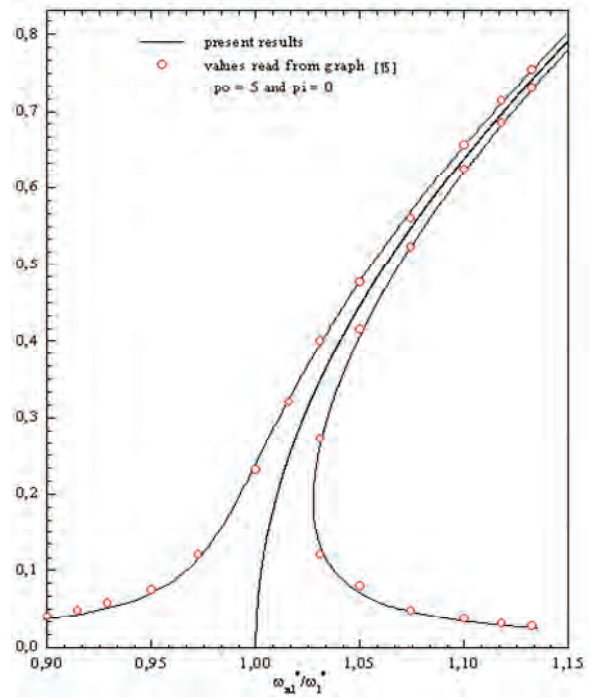


Figure 3 Comparison of the forced response curve obtained by present results and that Published in reference [15]. Plate 2 $(x^*, y^*) = (0.5, 0.5)$ and $F_d = 4 \text{ N/m}^2$

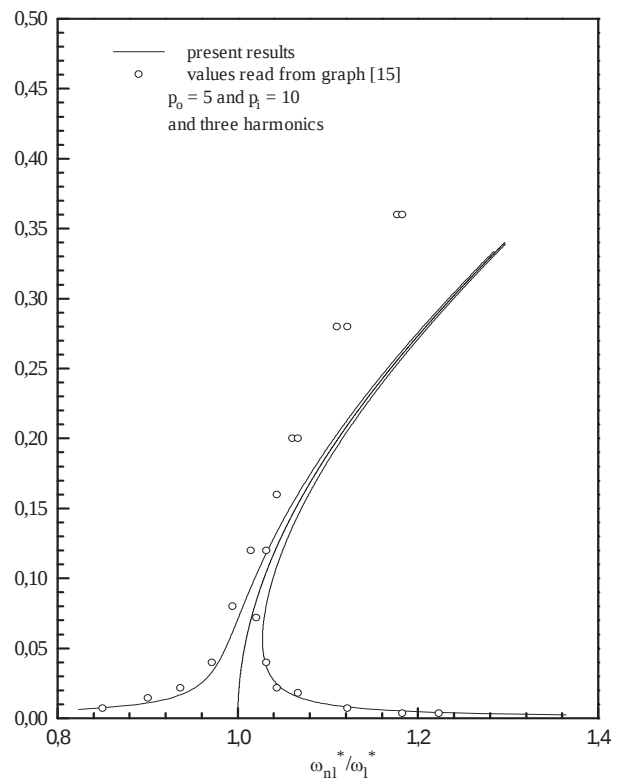


Figure 4. Comparison of the forced response curve obtained by present results and that published in reference [15]. Plate , $(x^*, y^*) = (0.75, 0.75)$ and $F_d = 4 \text{ N/m}^2$



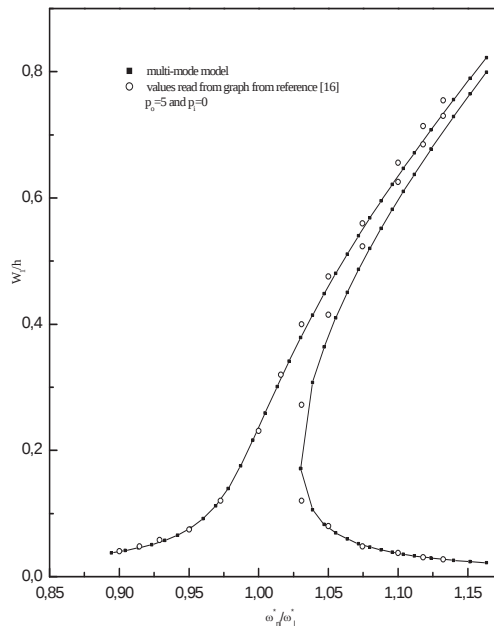


Figure 5. Comparison of the forced response curve obtained by multi-mode model and that published in reference [16]. $(x, y) = (0.5, 0.5)$ and $F_d = 4 \text{ N/m}^2$.

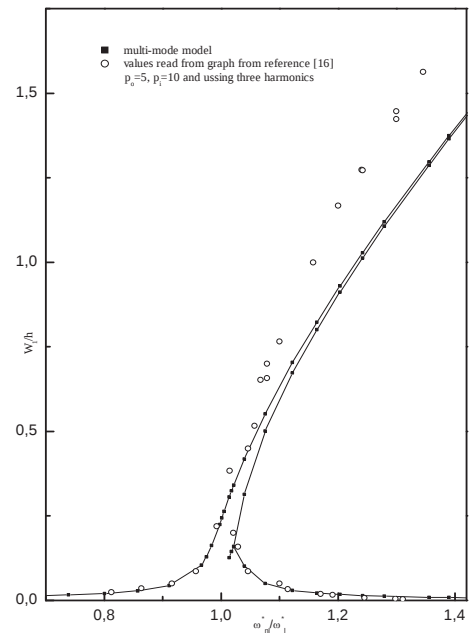


Figure 7. Comparison of the forced response curve obtained by multi-mode model and that published in reference [16]. $(x, y) = (0.5, 0.5)$ and $F_d = 4 \text{ N/m}^2$.

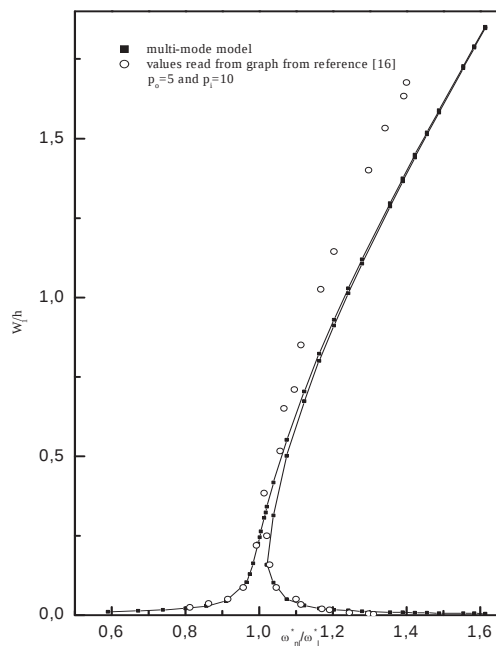


Figure 6. Comparison of the forced response curve obtained by multi-mode model and that published in reference [16]. $(x, y) = (0.5, 0.5)$ and $F_d = 4 \text{ N/m}^2$.

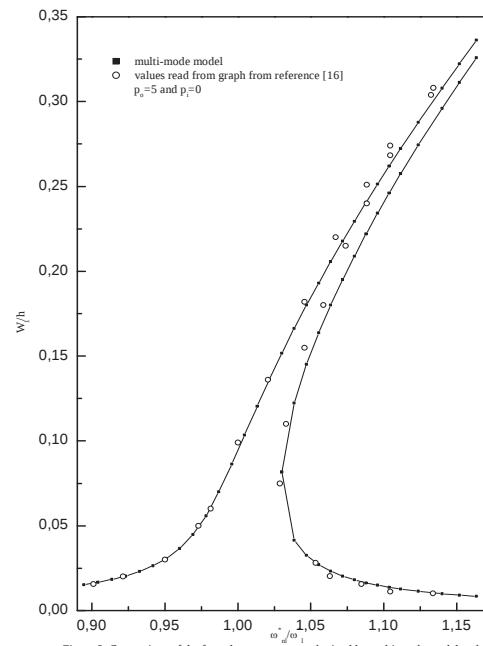


Figure 8. Comparison of the forced response curve obtained by multi-mode model and that published in reference [16]. $(x, y) = (0.75, 0.75)$ and $F_d = 4 \text{ N/m}^2$.



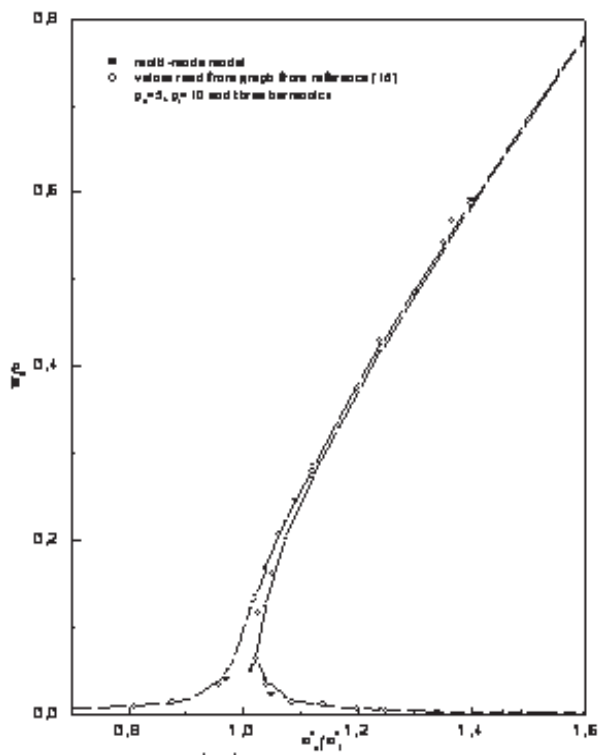


Figure 9: Comparison of the forced response curve obtained by multi-mode model and that published in reference [16]. $(x^*, y^*) = (0.75, 0.75)$ and $F_d = 4 \text{ N/m}^2$





A Robust Adaptive Control Algorithm for Sensorless Induction Motor Drives

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Abstract

In this paper, a robust adaptive control algorithm for speed sensorless induction motor drive is presented. The main feature of this control algorithm is that minimum synthesis is required to implement the proposed strategy. It is known as model reference adaptive controller (MRAC). The MRAC appeared to be robust in the fact of totally unknown plant dynamics, external disturbances and parameter variations. The induction motor stator resistance variations effect on the rotor speed are rejected by the MRAC controller due to its robustness. However, the performance of the Indirect Stator Field Oriented Control (ISFOC) are shading off. It is so because the reference commands computation of the ISFOC are based on the induction motor model, which includes that resistance. Therefore, a combination of three observers is used : a Luenberger observer to obtain the rotor flux and a two MRAS observers for rotor speed and stator resistance estimation. Digital simulation results are presented to highlight the improvement in performances of the overall proposed control scheme.

Keywords: *Induction motor, ISFOC, MRAC, sensorless rotor speed, stator resistance, MRAS and Luenberger observer.*

1 Introduction

In concert with the advances in factory automation, servo systems have become indispensable to many applications in industry. Process in the use of induction motors with high performance AC servo systems is remarkable. Previously, DC motors were used extensively for variable speed drive because of ease of control. However, AC motors have been controlled like DC motors using mainly a field oriented control approach [1]. Because of advances in power electronics and microprocessors, the induction motor drive

used in variable speed control has become more attractive. In view of the advances in power electronics and microprocessors, digitally controlled induction motor drives become increasingly popular. In many industrial drives, advanced digital control strategies for the field oriented control of induction motor with a conventional speed controller like such as the PI controller, have gained wide acceptance in high performance AC servo systems.

A high performance servo system must have good dynamic speed command tracking and load regulating responses, and the performance must be insensitive to system parameter variations. The conventional PI controller is easily implemented and highly effective if the load changes are small and the operating conditions do not take the system too far from the linear equilibrium point. However, in many applications as machine tools, the drive operates under a wide range of load changes and the system parameters vary substantially. In addition, because, it processes only one degree of freedom, the PI controller can't provide good tracking and regulating performance simultaneously. To overcome this drawback, the control algorithm should include a complicated computation process to eliminate the variations in load disturbance and system parameters and also obtain high performance AC servo systems. However, the control algorithms which are suitable to these systems have become increasingly more complicated, requiring extensive computations for real time implementation [2].

H_∞ loop shaping is used in the speed loop [3] and [4]. It remains robust against disturbances and load torque. However, this approach requires the knowledge of the limits of the disturbances. The sliding mode control is also used in the speed loop [5] but the most drawback of this type of controller is its higher switching frequency [6]. In recent years, speed controllers based on artificial intelligence techniques such as fuzzy logic [2], [7], [8] and [9] and neural network have [10] been widely proposed.



Since, these approaches don't require the knowledge of a mathematical machine model, the algorithm remains robust despite parameter deviations and noise measurement. However, the computation expenses and the requirement of expert knowledge for the system set up have seriously restricted their applications in practice [2].

Parameter variations represent a great problem not only for the speed controller, but also in the ISFOC decoupling of the induction motor [11]. As, the decoupling equations are deduced from the induction motor model which depends on the induction motor parameters. To overcome this problem, the identification and on line tuning of these parameters are strongly needed.

All high performance vector controlled induction motor drives depend on the quality of the signal measurement specially the rotor speed. As, it is used either in the speed control loop as in the field weakening operating region [12]. The principle of the field weakening operation consists of varying the flux reference in proportion to the reverse of the speed.

In a published research work [13], a direct adaptive control law is used in the speed loop. However, the effectiveness of the proposed method is restricted only to the load torque. Hence, robustness against motor parameter Variations as a major factor are not considered.

In this paper, a direct adaptive control based on the Model Reference Adaptive Control (MRAC) is performed for the speed control loop. The effectiveness of this algorithm is tested both with load torque and stator resistance variations. The ISFOC is used to assume the decoupling of the induction motor, the variation of stator resistance makes fails for this type of decoupling. To compensate the effect of the stator resistance variation, an algorithm based on the Model Reference Adaptive System MRAS principle is proposed to make an online identification and decoupling process tuning. The rotor speed is reconstructed by an MRAS speed observer. Digital simulation results are provided to show the improvement in performance of the proposed robust control of a sensorless IM drives. The paper discusses the Adaptive Control for Sensorless Induction Motor Drives. Section 2 presents stator flux orientation model. Section 3 illustrates Speed and stator resistance adaptive rotor flux observers. Section 4 introduces the adaptive scheme for stator resistance and rotor speed observation. Section 5 shows the design of an adaptive speed controller. Section 6 explains the simulation results and section 7 is the conclusions.

2 Stator flux orientation model

For ISFOC [1], [14] and [15], the stator flux vector is aligned with the d axis and setting the stator flux to be constant and equal to the rated flux, which means

$$\phi_{ds} = \phi_s \text{ and } \phi_{qs} = 0.$$

Assuming these conditions, the dynamic model of an induction motor can be represented according to the d and q axis components in a synchronous rotating frame as [11], [16], [17] and [18] :

$$v_{ds} = r_s i_{ds} + s \Phi_{ds} \quad (1)$$

$$v_{qs} = r_s i_{qs} + \omega_s \Phi_{ds} \quad (2)$$

$$i_{ds} = \frac{[1 + \tau_r s] \Phi_{ds} + L_s \tau_r \sigma i_{qs} \omega_r}{L_s [\tau_r \sigma s + 1]} \quad (3)$$

$$i_{qs} = \frac{T_e}{\Phi_{ds} p} \quad (4)$$

$$\omega_r = \frac{L_s [1 + \tau_r \sigma s]}{\tau_r [\Phi_{ds} - L_s \sigma i_{ds}]} i_{qs} \quad (5)$$

$$\omega_s = \omega_e + \omega_r \quad (6)$$

It can be seen if the stator flux is kept constant, the electromagnetic torque can be only controlled by the q axis current.

3 Speed and stator resistance adaptive rotor flux observers

Assuming that linear magnetic circuits, equal mutual inductances and neglecting iron losses, if the stator currents and the rotor flux are selected as the state variables, the state equations can be expressed in a stationary reference frame as [18] and [19]:

$$\begin{cases} \frac{d}{dt} \begin{bmatrix} \vec{i}_s \\ \vec{\phi}_r \end{bmatrix} = \begin{bmatrix} A_{11} & A_{12} \\ A_{21} & A_{22} \end{bmatrix} \begin{bmatrix} \vec{i}_s \\ \vec{\phi}_r \end{bmatrix} + \begin{bmatrix} B_1 \\ B_2 \end{bmatrix} \vec{v}_s, \\ \vec{i}_s = C \begin{bmatrix} \vec{i}_s \\ \vec{\phi}_r \end{bmatrix}. \end{cases} \quad (7)$$

$$A_{11} = -\frac{1}{\sigma L_s} \left(r_s + \frac{r_r M}{L_r} \right) I,$$

$$A_{12} = \frac{r_r M}{\sigma L_s L_r} I - \frac{M}{\sigma L_s L_r} \omega J,$$

$$A_{21} = \frac{r_r M}{L_r} I, A_{22} = -\frac{r_r}{L_r} I + \omega J,$$

$$B_1 = \frac{1}{\sigma L_s} I, B_2 = 0_2,$$

$$C = [I \quad 0_2], I = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}, J = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}$$

$$\sigma = 1 - \frac{M}{L_s L_r},$$

and



$$O_2 = \begin{bmatrix} 0 & 0 \\ 0 & 0 \end{bmatrix},$$

$\vec{i}_s = [i_{\alpha s} \ i_{\beta s}]^T$ is the stator currents vector,

$\vec{v}_s = [v_{\alpha s} \ v_{\beta s}]^T$ is the stator voltage vector,

$\vec{\phi}_r = [\phi_{\alpha r} \ \phi_{\beta r}]^T$ is the rotor flux vector.

r_s and r_r are the stator and rotor resistances respectively, L_s , L_r and M are the stator and rotor self and mutual inductances respectively, σ is the leakage inductance coefficient and ω is the rotor speed.

The electromagnetic torque developed by the motor is expressed in terms of rotor flux and stator currents as:

$$T_e = p * \vec{\phi}_r \otimes \vec{i}_s, \quad (8)$$

While the load torque acts as a disturbance according to the mechanical equation:

$$m_I \frac{d\omega}{dt} = T_e - T_L. \quad (9)$$

m_I is the rotor inertia, p the number of pole pairs and T_L the load torque.

The adaptive rotor flux observer, to get the estimated stator current and rotor flux using mainly the measured stator current $\vec{i}_s$ and the stator voltage $\vec{v}_s$, is described by the following state of equations:

$$\frac{d}{dt} \begin{bmatrix} \vec{i}_{se} \\ \vec{\phi}_{re} \end{bmatrix} = \begin{bmatrix} A_{11e} & A_{12e} \\ A_{21} & A_{22e} \end{bmatrix} \begin{bmatrix} \vec{i}_{se} \\ \vec{\phi}_{re} \end{bmatrix} + \begin{bmatrix} B_1 \\ B_2 \end{bmatrix} \vec{v}_s + Le(t)$$

$$\vec{i}_{se} = C \begin{bmatrix} \vec{i}_{se} \\ \vec{\phi}_{re} \end{bmatrix} \quad (10)$$

Where, "e", denotes the estimated values, L is the feedback gain matrix of the Luenberger observer and $e(t)$ is the stator current observation errors.

In this work, the matrix gain observer is obtained using the principle of the LRQ as proposed in [15], [16] and [18].

4 Adaptive scheme for stator resistance and rotor speed observation

The rotor speed and the stator resistance are reconstructed using the model reference adaptive system (MRAS) [18], [20] and [21]. The MRAS principle is based on the comparison of the outputs of two estimators. The first is independent of the observed variable named as model reference. The second is the adjustable one. The error between the two models feeds

an adaptive mechanism to turn out the observed variable.

In this work, the actual system is considered as the model reference and the observer is used as the adjustable one.

The stator resistance [22] and [23] and the rotor speed [24] are reconstructed by using the following adaptive mechanisms:

$$r_{se} = K_{p1} e_{is}^T i_{se} + K_{I1} \int_0^t e_{is}^T i_{se} dt \quad (11)$$

$$\omega_e = K_p e_{is} \phi_{re} + K_I \int_0^t e_{is} \phi_{re} dt \quad (12)$$

Where K_i , K_{i1} , K_p and K_{p1} are the integral and proportional constants respectively. As shown in figure 1, K_i , K_{i1} , K_p and K_{p1} are considered as the weights multiplying the product between current errors and the stator currents and the estimated rotor flux and its integral respectively. Therefore, the tracking performance of the speed estimation and the sensitivity to noise are depending on the proportional and integral coefficient gains. The integral gain K_i is chosen to be high for fast tracking of speed. While, a low proportional K_p gain is needed to attenuate high frequency signals denoted as noises [17].

MRAS representation structure for identifying the rotor speed and the stator resistance simultaneously is given in figure 1.

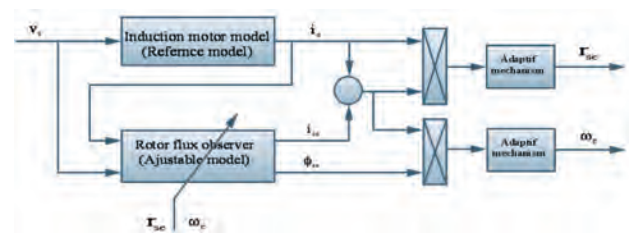


Figure 1: MRAS observed rotor speed and stator resistance.

5 Design of adaptive speed controller

The dependance of the induction motor parameters on the temperature is a serious hindrance in high performance control of the induction motor drives [5] and [25]. The temperature rise is caused by power losses in the motor [26] specially when the stator current exceeds its nameplate level. Under certain circumstances, such as overload or start up operations, the stator current can be much higher than the rated one, resulting in a significant increase of the stator resistance [27]. Compared to a cold motor, the induction motor parameters can increase by an order to 100% [25] and



[28]. The use of the classical controllers such as the proportionnel and integral controller is insufficient to provide good speed tracking performance. To overcome these problems, an adaptive controller based on the MRAC principle is proposed for the speed control law synthesis [13] and [29].

5.1 Discrete model reference adaptive control

we consider the single input single output discret linear time invariant plant of the first order system. It is described by the input - output pair $\{u(k), y_p(k)\}$ and defined with the transfer function $G(q^{-1})$ by:

$$G(q^{-1}) = \frac{y_p(k)}{u(k)} = \frac{b_1 q^{-1}}{1 + a_1 q^{-1}} \quad (13)$$

A model reference represents the desired behavior of the system. This model is described by the input - output pair $\{r(k), y_m(k)\}$. It is identified by the following transfer function:

$$G_m(q^{-1}) = \frac{y_m(k)}{r(k)} = \frac{b_m q^{-1}}{1 + a_m q^{-1}} \quad (14)$$

Where:

$r(k)$ is the reference input,

$y_m(k)$ is the output of the model reference.

The reference input $r(k)$ is bounded.

The objective of the numerical model reference control law $u(k)$ is to minimize the tracking error. This control law satisfies the following equation [30]:

$$\lim_{k \rightarrow \infty} e(k) = \lim_{k \rightarrow \infty} (y_p(k) - y_m(k)) = 0 \quad (15)$$

The minimization criterion is defined by:

$$J = \frac{1}{2} e(k) = \frac{1}{2} (y_p(k) - y_m(k)) \quad (16)$$

We designate by θ the vector of the controller parameters. To update these parameters, we adopte the gradient algorithm. It is defined by:

$$\theta(k+1) = \theta(k) - \alpha \frac{\partial e(k)}{\partial \theta(k)} \quad (17)$$

With:

α is the adaptation factor.

The control law which minimizes the criterion J is given by [31]:

$$u(k) = t_o(k)r(k) - s_o(k)y_p(k) \quad (18)$$

Where $t_o(k)$ and $s_o(k)$ are the components of the controller vector.

$$\theta(k) = \begin{bmatrix} t_o(k) & s_o(k) \end{bmatrix}^T$$

By replacing each element with its expression in (17), we obtain:

$$\begin{pmatrix} t_o(k+1) \\ s_o(k+1) \end{pmatrix} = \begin{pmatrix} t_o(k) \\ s_o(k) \end{pmatrix} - \alpha \begin{pmatrix} \frac{\partial e(k)}{\partial t_o(k)} \\ \frac{\partial e(k)}{\partial s_o(k)} \end{pmatrix} \quad (19)$$

After replacing $u(k)$ with its expression in (13), the model output can be written in the following form:

$$y_p(k) = \frac{b_1 t_o(k) q^{-1}}{1 + (a_1 + b_1 s_o(k)) q^{-1}} r(k) \quad (20)$$

To formulate the error, we subtract the reference output $y_p(k)$ of (20) to model output. Subtracting $y_p(k)$ from $y_m(k)$ of (14), we obtain:

$$e(k) = \left[\frac{b_1 t_o(k) q^{-1}}{1 + (a_1 + b_1 s_o(k)) q^{-1}} - \frac{b_m q^{-1}}{1 + a_m q^{-1}} \right] r(k) \quad (21)$$

The differentiation of the error through $\theta(k)$ can be written in the following form:

$$\frac{\partial e(k)}{\partial \theta(k)} = \begin{bmatrix} \frac{b_1 q^{-1}}{1 + (a_1 + b_1 s_o(k)) q^{-1}} r(k) \\ -\frac{b_1 q^{-1}}{1 + (a_1 + b_1 s_o(k)) q^{-1}} y_p(k) \end{bmatrix} \quad (22)$$

In order to obtain a dynamic system equivalent to the reference model, we impose the characteristic equation. Therefore,

$$\frac{\partial e(k)}{\partial \theta(k)} = \begin{bmatrix} \frac{b_1 q^{-1}}{1 + a_m q^{-1}} r(k) \\ -\frac{b_1 q^{-1}}{1 + a_m q^{-1}} y_p(k) \end{bmatrix} \quad (23)$$

Replacing $\frac{\partial e(k)}{\partial \theta(k)}$ with its expression in (19), we obtain:

$$\begin{cases} t_o(k+1) = t_o(k) - \alpha \frac{b_1 q^{-1}}{1 + a_m q^{-1}} r(k) \\ s_o(k+1) = s_o(k) + \alpha \frac{b_1 q^{-1}}{1 + a_m q^{-1}} y_p(k) \end{cases} \quad (24)$$

The developpement of these equations yields:

$$t_o(k) = (1 - a_m) t_o(k-1) + a_m t_o(k-2) - \alpha e(k) b_1 r(k) \quad (25)$$

$$s_o(k) = (1 - a_m) s_o(k-1) + a_m s_o(k-2) + \alpha e(k) b_1 y_p(k) \quad (26)$$

The structure of the direct adaptive control with the MRAC principle is shown in figure 2.

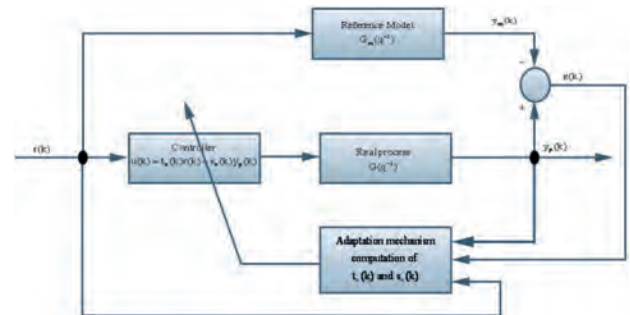


Figure 2: The basic structure of the Model Reference Adaptive Control.



5.2 MRAC for induction motor drive

If we suppose that the load torque is as a disturbance, then using the mechanical equation, we obtain:

$$\frac{\omega(s)}{T_e(s)} = \frac{k}{1 + \tau s} \quad (27)$$

where $k = \frac{f}{m_I}$ and $\tau = \frac{1}{m_I}$

The discrete transfert function can be written as:

$$\frac{\omega(k)}{T_e(k)} = \frac{b_1 q^{-1}}{1 + a_1 q^{-1}} \quad (28)$$

The model reference must be chosen for the same plant as the real one. Therefore,

$$\frac{\omega_m(k)}{\omega_{ref}(k)} = \frac{b_m q^{-1}}{1 + a_m q^{-1}} \quad (29)$$

Where b_m and a_m are the model reference parameters. The command law of the model reference is given by the recurrent equations:

$$u(k) = t_0(k) \omega_{ref}(k) - s_0(k) \omega_e(k) \quad (30)$$

$t_0(k)$ and $s_0(k)$ are updated from the two following recurrent equations:

$$\begin{aligned} t_0(k) &= (1 - a_m) t_0(k-1) + a_m t_0(k-2) - a e(k) \\ b_1 \omega_{ref}(k-1) \end{aligned} \quad (31)$$

$$\begin{aligned} s_0(k) &= (1 - a_m) s_0(k-1) + a_m s_0(k-2) + a e(k) \\ b_1 \omega_{ref}(k-1) \end{aligned} \quad (32)$$

The proposed configuration of the robust adaptive control law is detailed in figure 3.

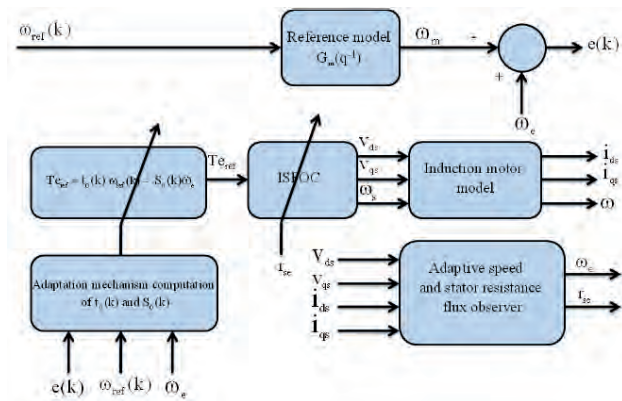


Figure 3: Configuration of the robust adaptive sensorless control system.

6 Simulation results

The dynamic behavior of the system was investigated using computer simulations with Matlab / Simulink

software.

We have tested the robust adaptive control algorithm for sensorless induction motor at different speed targets under no load and with load applied. The effects of stator resistance variations is also studied.

The results obtained at variable step speed targets under no load are reported in figures 4 to 7.

It can be clearly seen that the d and q axis stator flux converge to the nominal value and to zero respectively. This indicates that the decoupling of the induction motor is established and ensured.

The system and the model reference responses at no load operation are given in figure 6. It can be seen that a good agreement between system and model reference responses is obtained. The feedforward and feedback gains are shown in figures 4 and 5 respectively. It is noticed that when an error between the system response and the model reference one occurs, the feedforward and feedback gains react to minimize this error. They converge to a fixed values when the error between the system and the model reference responses becomes zero.

To investigate the behavior of the proposed sensorless control drives against external disturbances, a load torque was added. The results obtained under load applied are shown in figures 8 and 9. The obtained results show that an error between the system response and model reference one occurs. It converges to zero thanks to the on line adaptation scheme of the controller parameters.

The robustness of the proposed sensorless control drives is tested according to the stator resistance variations. The obtained results are given in figures 10 and 11. It is shown that the system response converge to the model reference response. An error between the system and the model responses occurs when stator resistance varies. It is compensated thanks to the adaptive identification of the controller parameters.

As it is seen in figure 11, a decoupling error occurs at a time of the stator resistance variation. But, this error is maintained and there is no compensation. This is because of the ISFOC commande equations which are established from the nominal value of the stator resistance.

To overcome the sensivity of the vector control algorithm (ISFOC) against stator resistance variations, an identification and on line adjustment of the stator resistance in ISFOC algorithm is necessary to maintain the decoupling of the induction motor. We have used the MRAS principle to observe the stator resistance value. The obtained results with on line adjustment of the stator resistance are reported in figure 13 to 15. The system response converge to the model reference one. An error occurs at a time of the stator resistance variation. It converges to zero thanks to the adaptive control algorithm. The target, observed and actual components of the stator flux are shown in figure 14. The d and q axis stator flux converge to the nominal



value and to zero respectively despite of the stator resistance variations. This is because of the on line identification and tuning of stator resistance in the ISFOC algorithm.

The observed and the actual stator resistance are given in figure 15. The results show that the observed stator resistance follow exactly the actual stator resistance, indicating the validity of the proposed algorithm.

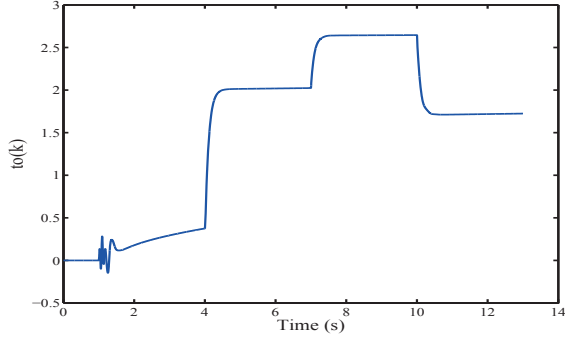


Figure 4: Evolution of $to(k)$ at variable target rotor speeds under no load.

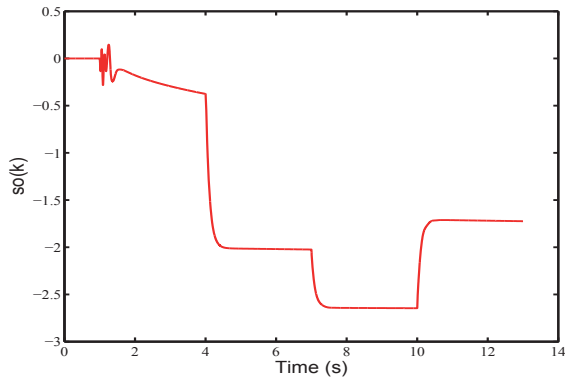


Figure 5: Evolution of $so(k)$ at variable target rotor speeds under no load.

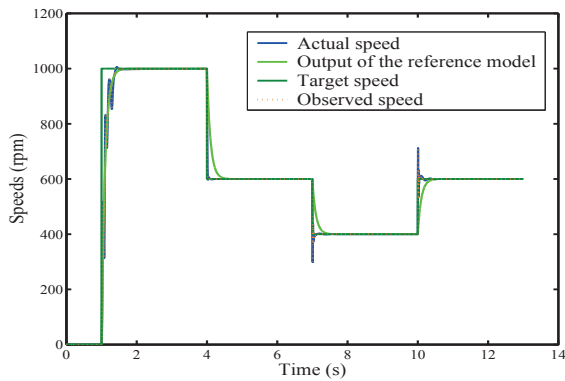


Figure 6: Actual, MRAS observed and output of the reference model at variable target rotor speeds under no load.

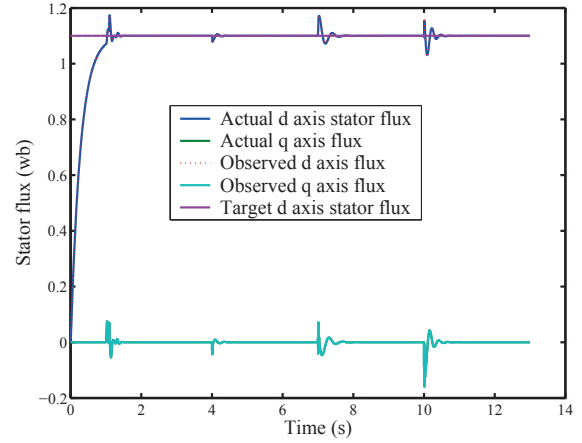


Figure 7: Actual, observed and target d - q components of stator flux at variable target rotor speeds under no load.

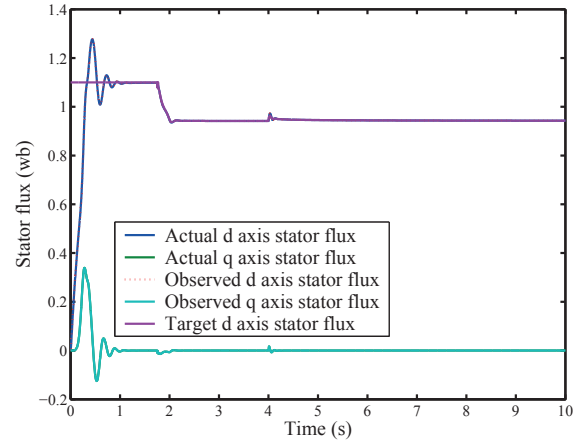


Figure 8: Actual, MRAS observed and output of the reference model at variable target rotor speeds under load.

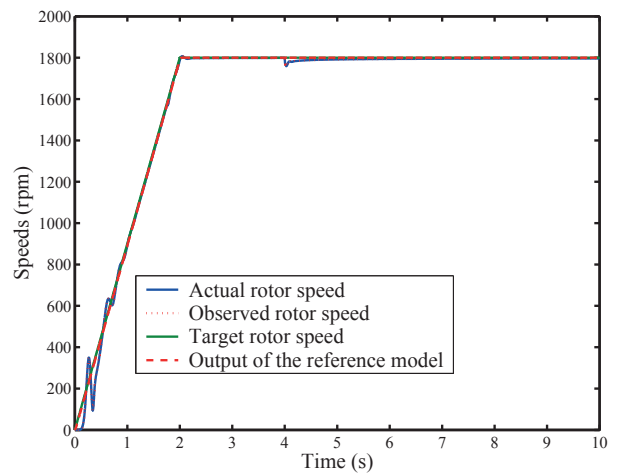


Figure 9: Real, observed and target d - q components of stator flux at variable target rotor speeds under load.



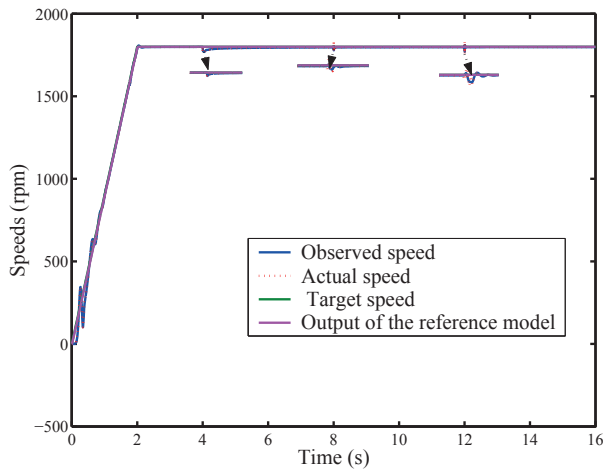


Figure 10: Actual, MRAS observed and target rotor speeds without on line adjustment of stator resistance.

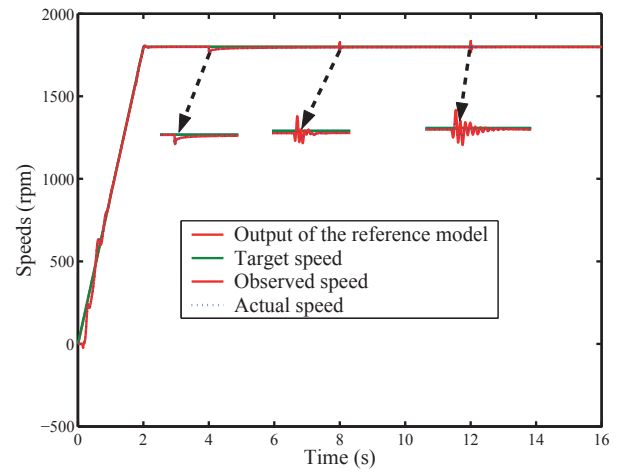


Figure 13: Actual, MRAS observed and target rotor speeds with on line adjustment of stator resistance.

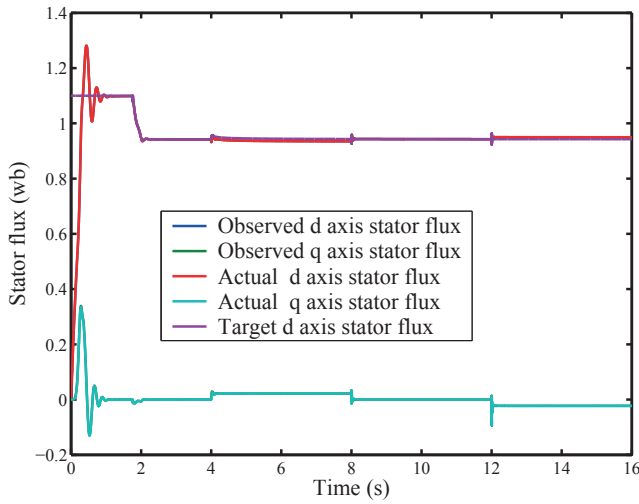


Figure 11: Actual, observed and target d - q components of stator flux without on line adjustment of stator resistance.

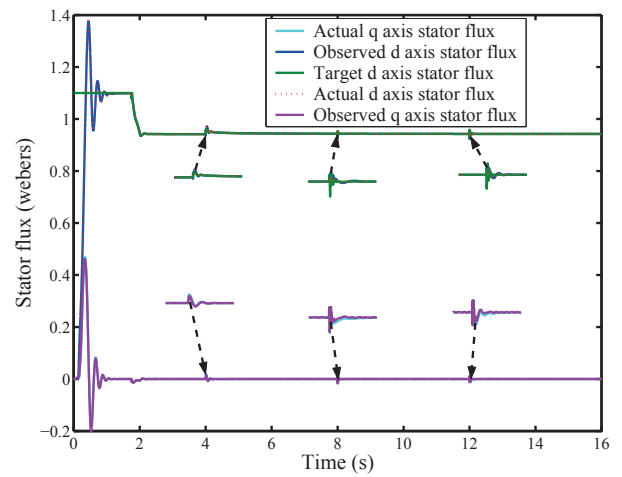


Figure 14: Actual, observed and target d - q components of stator flux with on line adjustment of stator resistance.

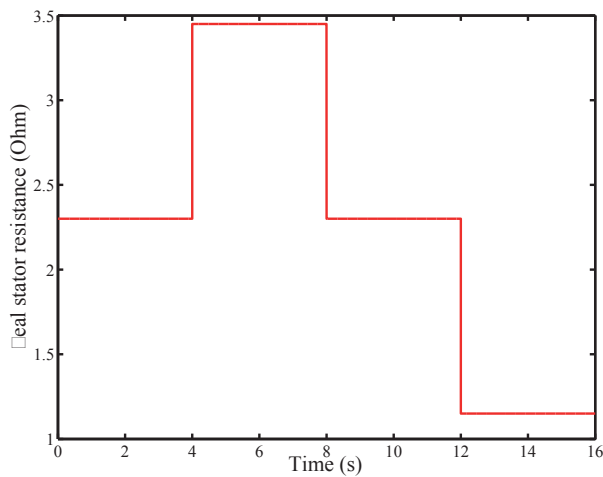


Figure 12: Actual stator resistance trajectory variation.

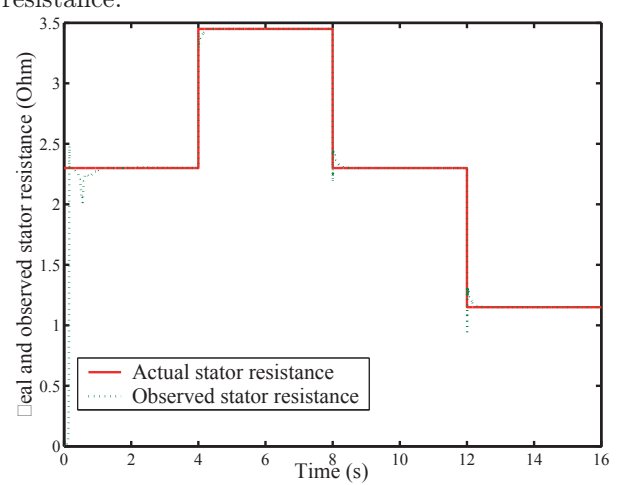


Figure 15: Actual and MRAS observed stator resistance.



7 Conclusion

In this paper, we have presented the synthesis of a robust adaptive control algorithm for sensorless induction motor drives. The proposed method has been applied to an indirect stator field oriented induction motor drive system. The adaptive with control algorithm for the rotor speed shows good performance under no load, load applied and against stator resistance variations. The influence of the stator resistance variation on the ISFOC algorithm was eliminated by the proposed algorithm for the identification and on line ajustement of the stator resistance. The validity of the robust adaptive control algorithm for sensorless induction motor drives has been verified by simulation under Matlab/Simulink environment. Practical implementation of the proposed method is a subject of future follow up research work.

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Study of a Project for Producing of Wind Energy: Design, Monitoring and Impact on the Electricity Grid

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Abstract

This paper will provide an effective help to all those who make decisions regarding the planning and implementation of projects wind Energies. We used the software Alwin for sizing a wind farm of 140 MW installed in Melloussa cite, 25 Km from Tangier, in north Morocco. We analyzed the collected wind data during one year, to assess wind power potential, the prediction of produced electric power in the site based on a judicious choice of Wind Turbine Generator, as well as, the automatic determination of direction of the wind on the site and analysis of wind turbulence. Thus, the integration into the grid provides several benefits.

However, this type of production which has an impact on the electric network and its connection is subject to technical constraints. We will study the various generators used to produce electric power from wind energy and choose the one with a good price per kilowatt hour, then to study the feasibility of integration of this wind farm in the electric grid. And finally, we treated the various principles, concepts used in data transmission and remote monitoring of wind farms, and we proposed industrial solutions of telemanagement which can be applied in remote wind farm in the site.

Keywords: *Wind data, Wind park, ALWIN software, Aerogenerators, Flicker, harmonics, Telemanagement.*

1. Introduction and the general objective

To optimize the exploitation of the wind farms, some measures must be taken into consideration. Indeed, a bad choice of certain parameters could harm a wind turbine establishment. The reason way a pre-study phase appears of extreme importance to identify the conditions and constraints to be taken into account in the realization of a wind project [1]-[2].

We must therefore proceed with a study and analysis of the climatology of the site considered, since a good knowledge of the site allows a better exploitation and production of electrical energy. However, the coupling of the Wind turbines at national grid has an impact on power quality and dependability. As a result, the connection is subject to certain technics and requires some adjustments in the network to ensure a proper functioning of this latter.

Thus, to better optimize and secure wind farms, a control system and remote monitoring must be established. This latter must be designed to allow automatic operation of wind farms.

And therefore, the general objective of this work is to help develop indigenous energy sources to supply electricity and profit from great potentials of wind power in Morocco as long it's a clean, safe and a free source of energy.

This paper is organized as follows: section 2 explains the Study feasibility of dimension of wind farm via Alwin software, and at section 3, we present the results of dimensioning a wind farm of Tangier, in section 4 we introduce Impact of the production of wind farm on the national network, we describe in section 5 the Concept of



remote wind farm. In the end, section 6 provides conclusions and directions for future research.

2. Dimension of wind farm via software

Alwin

2.1. Sizing techniques: [3]-[4]-[5]-[6]

2.1.1. Wind power:

$$P = \left(\frac{1}{2}\right) \rho S V^3 \quad (1)$$

$$V_{moy} = \sum V_i f_i \quad (2)$$

- P: wind power (kw),
 ρ : the density of air (Kg/m³)
 V: the average wind speed in (m / s)
 S: the area swept by the wind turbine in (m²)
 V_{moy}: the average wind speed in (m / s)
 f_i: Frequency in% of the wind speed V_i in (m/s) to class i.
 i: class of wind speed
 V_i: the wind speed in (m / s) at f_i frequency

2.1.2. Energy provided by a wind turbine:

$$E = T_h \times \sum P_i(V) f_i(V) \quad (3)$$

With:

E:	energy produced kWh.
T _h :	time in hours
P _i (V):	power (KW) given by the wind speed V _i
f _i (V):	frequency of wind speed V _i

2.1.3. Capacity Factor

The capacity factor assesses the degree of utilization of the turbine, it is defined as follows:

$$C(\%) = (100 \times P_{moy}) / P_{max} \quad (4)$$

These two powers P_{moy} and P_{max} are chosen for the wind turbine. As the capacity factor is high, it is certainly good choice of the wind turbine.

- P_{max}: maximum power delivered by aerogenerator
 P_{moy}: average power delivered by aerogenerator

2.1.4. The total number of turbines to be placed on the site [1] Conditions to be attained:

$$\begin{aligned} (N1 + 1) * 10H &< I \\ (N2 + 1) * 3D &< L. \\ N &= N2 * N1 \end{aligned} \quad (5)$$

Key:

- I: dimension of the field perpendicular to the prevailing wind direction.
 L: size of land parallel to the prevailing wind direction
 D: rotor diameter of the machine
 H: height of the tower
 N1: number of turbines per row

N2: number of row of turbines

N: total number of turbine to be placed on the site

Mathematical modelling of frequency distribution of wind: distribution WEIBULL:

As it is difficult to manipulate all of data on a frequency distribution of wind, it is more convenient for theoretical considerations to model the frequency histogram of wind speeds by a mathematical function that continues through discrete values table. We can therefore choose the model WEIBULL [3]-[4]-[5]. Indeed, for periods ranging from weeks to a year, the WEIBULL function reasonably represents the observed rates. There is a probability density function in the following form:

$$P(V) = (K/C)(V/C)^{k-1} \exp(-(V/C)^k) \quad (6)$$

With:

- P (V): probability density of the velocity V
 K: form factor of the curve (dimensionless)
 C: scale factor of the curve (meter / second)

The average wind speed can be found by integrating the probability density is therefore the formula (6):

$$V_{moy} = \int V P(V) dv \quad (7)$$

$$\text{Or } V_{moy} = C \Gamma(1 + 1/k) \quad (8)$$

Where $\Gamma(x)$ is the mathematical function Gamma, defined by:

$$\Gamma(x) = \int_0^{\infty} e^{-t} t^{x-1} dt \quad (9)$$

We can deduce other important parameters such as the average power:

$$P_{moy} = 0.5 \rho C^3 \Gamma(1 + 3/k) \quad (10)$$

Thus, the distribution WEIBULL can facilitate many calculations that are necessary for the analysis of wind data.

2.2.1. Determination of WEIBULL parameters [4]-[5]:

Now, we must focus our attention on the parameters K and C, as they are the key to using the WEIBULL distribution. These parameters define the shape and scale of the curve WEIBULL.

The form factor, K (dimensionless), expressed in the form of WEIBULL curve: A high value of K implies a narrow distribution, with winds concentrating around a value, while a low value of K leads to widely winds dispersed. The factor C (m/s), determines the position of the curve. Generally, its value is high for windy sites and low for quiet sites. Various methods can be applied to determine K and C from a given wind distribution.



$$C = 1.125 V_{moy} / (1 - B)$$

$$K = 1 + 0.483 (V_{moy} - 2)^{0.51}$$

$$B = 1 - 0.81 (V_{moy} - 1)^{0.089} \quad (11)$$

This probability density function is called Rayleigh distribution when $K = 2$

2.2.2. Changes in wind direction:

We can determine the dominant wind direction. The alignment of the turbine is intended to capture the maximum energy available in the wind. Thus, in so far as the ground permitted, wind turbines should be aligned perpendicular to the predominant directions of the wind. In general, the wind direction is presented as a wind rose.

The parameters in the paragraph above "sizing techniques" describe the characteristics of wind in any site.

Winds are generally characterized by:

- The average speeds (hourly, daily, monthly and annually).
- The predominant directions.
- The frequency of each of the speeds and directions.
- Diurnal Pattern / Duration of calms / calm analysis.
- A good knowledge of these characteristics is important in so far as they are involved in:
 - o The choice of wind turbines and the evaluation of their performance;
 - o The implantation (Proper orientation of wind).

3. Presentation of results of dimensioning [5]- [7]

3.1. Wind Potential of Tangier's site

We processed and analyzed the data of station in Tangier of 369 days for an assessment of the wind potential using the menu under "Annual Review". We found the frequency distribution of wind speed as follows:

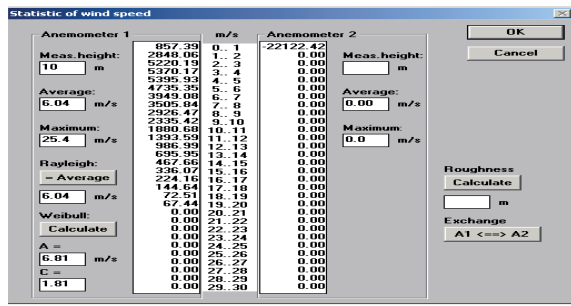


Figure 1. a: Statistic of Wind potential of Tangier's site

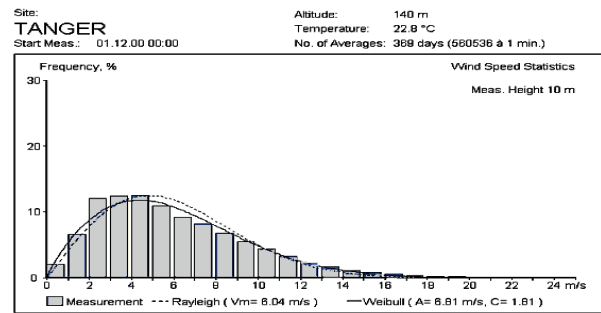


Figure 1.b: Wind potential of Tangier's site

The annual review of the frequency distribution of the wind speed gives the following information: The average speed of the site is: 6.04m/s. The maximum speed of the site is: 25.4m/s. Speeds predominant site varies between 2.5m/s and 11 m/s.

3.2. Daily variation / calm analysis [5]-[7] :

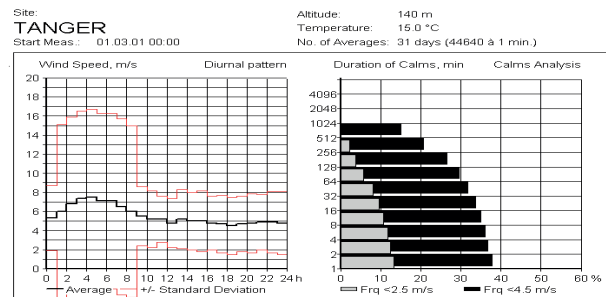


Figure 2. Daily variation / curve calm.

On the one Hand, the analysis of wind turbulence in the site through the curve of the daily model allows us if there is a continuity of service (Permanent Power) at the station. On the other hand, the curve calm analysis of the station in the form of sticks shown in Figure 2 shows the cumulative frequency percentage for speeds below 2.5 m/s and 4.5 m/s and remain for some time calm. The determination of the frequency of calm is needed to determine the speed boot of aerogenerator which will be used for energy production. Also, we need the frequency of calm for determining the storage capacity of batteries included in the electrification systems (independent), as well to ensure continuity servicing production of energy.

3.3. Predicted annual energy produced by the wind farm based on the appropriate choice of wind turbine [5]- [7]

The submenu "Prediction of Electric Power" is designed to show the user the power measured in MWh, the average power of the wind turbine selected and the capacity factor, which assesses the degree of utilization of wind turbine,



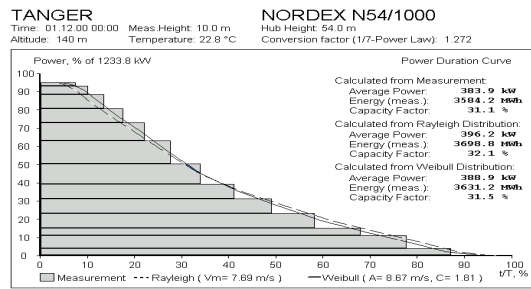


Figure 3.a: Curve prediction of energy.
Wind Energy Converter: NORDEX N54/1000
Catalog: DEVI8503 CAT
Hub Height: 54.0 m
Control: STALL

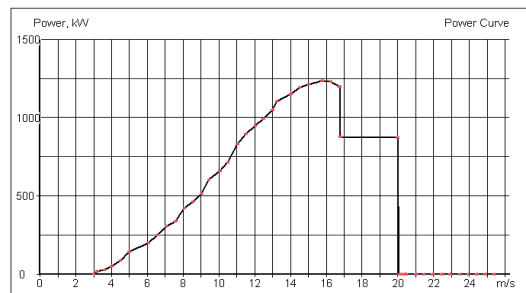


Figure 3. b: the machine power curve characteristic

- The choice "NORDEX N54/1000," $P_n = 1\text{ MW}$ can produce an energy of 3.69 GWh annually by each turbine.

Knowing that the installed capacity of 140MW, or 140 Wind turbines give the annual energy production of 516.6 GWh in the site.

3.4. Result of wind Direction [5]-[7]

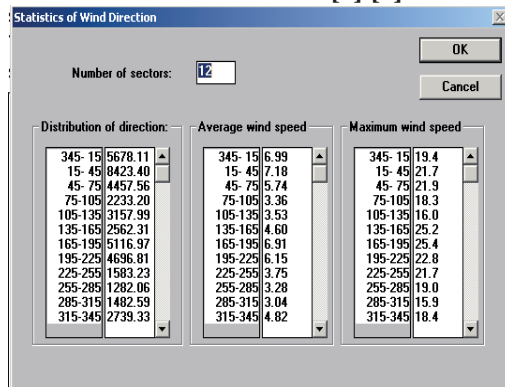


Figure 4.a: Wind direction Statistic: Tangier's site.

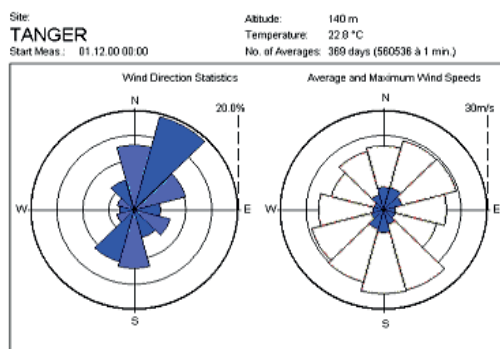


Figure 4.b: Wind rose Tangier's site

The annual review can opt for the north-east, specifically an inclination with respect to North of a

varying geometric angle between the first and the third a sector (between 15 degrees and 45 degrees).

4. The Impact of the wind production park on the national network

4.1. Phenomena of flicker and harmonics [8]-[9]

The flicker [8]

The flicker is a visual sensation produced by rapid changes of an effective tension.

The main cause of flicker emission by the wind turbines is a turbulent wind.

- Other phenomena can cause flicker as:

- * The shadow effect of the tower.
- * The blades error pitch.
- * The rotor orientation error.

The Harmonics

The harmonics are introduced by elements of electronics power, especially the frequency converters, the thyristors controllers and capacitors.

4.2. The choice of most appropriate turbines

4.2.1. The different technologies of wind turbines (aerogenerator)

Table 1 summarizes the different technologies of wind turbines, we opted for the DFIG for the following reasons:

- Possibility to regulate the tension in connection point where it is injected (very small flicker) [10]
- Asynchronous double fed machine gathers the advantages of asynchronous and synchronous machines.
- Rugged and allows the inspection of the reactive energy.
- Costly compared to the asynchronous machine, but it is less than the synchronous.

A kilowatt hour price much better the than synchronous machine under the same regime of wind [10].

4.2.2. The choice of the machine double fed induction asynchronous generator (DFIG)

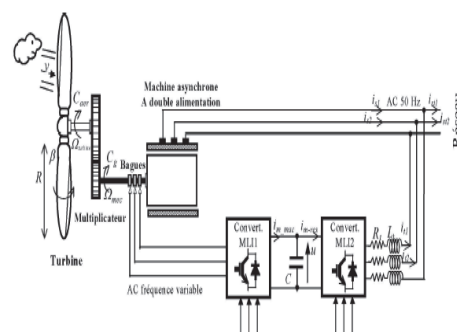


Figure 5. The double-fed asynchronous machine



Table 1. Summary of the different technologies [6]-[8]-[9] of wind turbines

Speed control	Type of generator	Reactive power	Disturbance created in the network	remark
Fixed speed With gearbox	Synchronous	Can be controlled by the current excitation	Mechanical oscillation because the generator is connected to the main frequency (flicker).	More used in large units
	Asynchronous	Requires a correction device $\cos(\phi)$	Requires a soft start device. Flicker generation	Use in the past in small units.
A speed adaptable within limits with gearbox	Asynchronous adaptation of rotor slipping resistance	Requires a correction device $\cos(\phi)$	-Soft start device -Possible Harmonic by adjusting the slipping.	Poor performance.
Variable Speed Rotor With gearbox	Asynchronous dual power	Power factor control by frequency converter rotor.	Flicker reduced but harmonics Problem	Solution technically interesting but we must see the investment cost
	Asynchronous dual rotor winding	Requires a correction device $\cos(\phi)$	Requires a system of soft start device	Solution technically interesting but we must see the cost of investment
Rotor without Variable speed gearbox	Multiple Synchronous	Can be controlled by the current excitation	Little disruption on networks.	Solution technically interesting but greater cost compared to the DFIG.

4.3. The flicker emission calculation

4.3.1. In continuous operation [9]

$$P_{st} = P_{lt} = N^{0.5} \cdot C(\psi) \cdot \frac{S_n}{S_{cc}} \quad (9)$$

$C(\psi)$: coefficient flicker of 10.4 (value given by the standard 61400-21).

S_n is the rated apparent power of the wind turbine.

S_{cc} is the apparent power of short-circuit at point of common coupling (PCC).

4.3.2. During operations commutation

$$P_{st} = \frac{18}{S_{cc}} \left[N \cdot N_{10} (Kf(\psi) \cdot S_n)^{3.2} \right]^{0.31}$$

$$P_{lt} = \frac{8}{S_{cc}} \left[N \cdot N_{120} (kf(\psi) S_n)^{3.2} \right]^{0.31} \quad (10)$$

$Kf(\psi)$ Factor: is the flicker of the wind for the phase angle impedance given network

According to international standard: IEC (International electrotechnical committee 61400-21 of measurement and evaluation of power quality characteristics of grid connected wind turbines. IEC 61400-21 involves: $N_{10} = 10$ and $N_{120} = 120$. Pst: Compatibility levels at short time for flicker networks Low Voltage and Medium Voltage. Plt Compatibility levels at long time for flicker networks Low Voltage and Medium Voltage. To evaluate the worst flicker, we take $Kf(\psi) = 1$.
4.3.3. Presenting the results of flicker calculations at Pcc [9]-[11]

Table 2. The results of the flicker calculations at Pcc

Active power (KW)	Power (KVA)	Number of unity	Post connectin g	Scmin (MVA)	Emission of flicker		
					continious	During switching operations	
					Pst=Plt	Pst	Plt
1000	1111.11	140	Wind Park	2779	0.0491	0.0608	0.0584
			Post of Melloussa	3379	0.0404	0.05	0.0480
			Post of Tanger	3199	0.0427	0.0528	0.0507

The flicker factors below 1, which complies with the standard, have been obtained.

4.4. Harmonics Calculation at PCC:

The harmonic current order h at the PCC due to N wind turbine [9]:

$$I_h = N \frac{1}{\alpha} \frac{I_{hi}}{n} \quad (11)$$

Key:

I_{hi} : the harmonic distortion of current order h due to a single wind turbine.

n : the report of a transformer wind turbine.

The coefficient α according to the order of harmonics.

The rates of harmonic tension due to 140 wind turbines at bus bar MT are in table 3.

Table 3. The rates of harmonic tension due to 140 wind turbines

	harmonic pair	odd harmonics	total harmonic
Value relative to 22kV	0,28%	1,19%	1,22%

The ONE (National Office of Electricity) requires the following rates of harmonics [11]:



Table 4. The rates of harmonics required by ONE

kind of harmonic	Pair	odd
Rate of harmonic	0,6%	1,5%

- The rate of overall harmonic is therefore well below the 1.6% imposed by ONE [11].
- The rate of harmonic issued by all the wind turbines is not embarrassing for the network and does not exceed the norms (limits prescriptives).

5. Concept of remote wind park

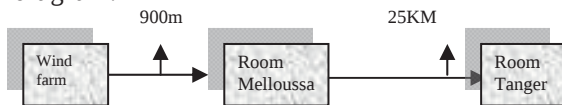
5.1. Purpose of the remote [12]

The remote is a powerful tool in management and control. It improves the speed of intervention of maintenance services and increases the performance and availability of machines. It is transmitted on a physical data and messages of breakdowns to a room monitoring.

5.2. Solutions proposed for remote for the park of Tangier

5.2.1. Problem Analysis

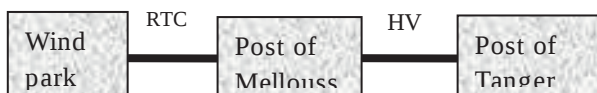
We want to monitor the wind farm in two different locations: the village post of Melloussa and the post of the city of Tangier as shown in the following diagram:



5.2.2 Data Transmission

Possible solutions

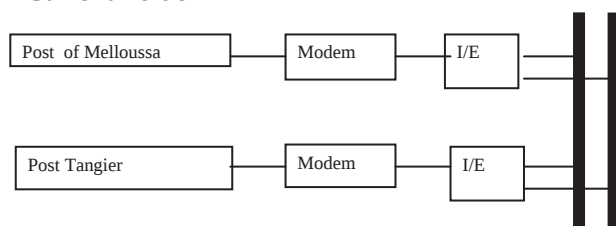
Each compact station at the foot wind turbine has a telephone line from the central Melloussa. The latter is linked to the central Tanger by a high voltage (225kV), we adopted the following solution:



- RTC (*Réseau Téléphone Comuté: phone network commuter*) and transmission power line on the high voltage line

Forwarded by RTC and current holders on the high-voltage line:

- Current holder link



I/E : injector/extractor.



- Reasons for the choice:

It's reliable, economic (cable transmission is the same cable HV) and already operated by ONE (National Office of Electricity) in Morocco at the factory IDRIS 1.

5.3. Control and monitoring of wind turbines [5]

5.3.1. Solution based on PLC (Processing logic control)

The configuration of the wind farm permits to consider only one solution if it is based controller, for each turbine, we will install a mini-controller.

5.3.2. Reasons for choice:

Maximum automation for a minimal investment, using a stand-alone and far simplicity of installation, programming and manipulation [13]-[14]-[15].

5.3.3. Architecture of the solution to be applied in the park of Tangier

At the foot of each turbine mini-PLC is installed.

The CPU will be installed in the room Melloussa and manages all micro-controllers before contacting the supervisor PC installed in Tanger.

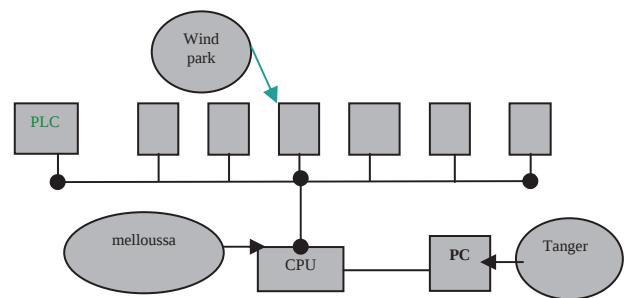


Figure 6. Architecture of the solution to be applied in the park of Tangier

CPU: Central programmable unity.

PLC: Processing logic control (PLC micro)

The micro-robot (SIMATIC S7-200) meets the desired functionality [9]. Already operated by the wind Park of Tetouan in Morocco, the only drawback its high cost.

5.3.4. Solution based on a microcontroller

The number of entry / exit and the size of the programs at each turbine are not important, so a microcontroller based solution seems very reasonable, because it replaces the role of the controller with a lower price.

5.3.5. Choice of network topology [13]-[14]-[15]:

In the star topology, all stations of wind turbines are connected directly to a server or a hub that is the central node through which all transmissions pass. This topology allows to easily add equipment (one cable per device in the limit of the server capacity). Network management is facilitated by the fact that the equipments are directly searchable by the server and all transmissions spend (centralization of software). Moreover, a failure of terminal equipment



does not affect the operation of the rest of the network. However, it can lead to significant lengths of cable. It is more economical than the parallel bus. And given the network structure, this solution is very appropriate for our application.

In the case of topology bus, if the device is the optical type, the adding operation is more difficult because it requires cutting the fiber to the location of the connection and that the stations located downstream report to the server will be blocked.

NOTE. - The connection between the PC Melloussa and wind turbines may also be performed by a RS 485:

There are two types of interface circuits: asymmetrical and differential.

5.3.6. The asymmetric-RS232 interface:

Currently the RS232 interface is regularly used for almost devices on the PC.

This standard requires a length of 15 m and has a low immunity to noise [13]-[14]-[15].

5.3.7. The differential interface-RS485:

These standards provide a universal solution for transmission over long distances. It may be taken as the solution for our case [1].

5.3.8. Reason for the choice of RS485:

It allows: a transmission distance up to 1200m, and a good immunity to electromagnetic interference.

As in our case, the distance is 900m and the site has electromagnetic interference, this solution is adequate.

5.3.8. Communication protocol [13]-[15]:

The link level protocol defines the set of rules on the meaning of data in the order in which they are transmitted. It prevents access conflicts.

We can distinguish two communication protocols in our case: Sequence of reading the information, and the other of writing

5.3.9. Data processing at Melloussa:

Once the data is transmitted by the PC, the database available at the central unit, will be treated later by using the software. Example of software Alwin studied in a first part of this work, with its system of data acquisition [16].

5.4. Industrial solution proposed for remote and monitoring the wind farm of Tangier:

According to industry solutions that we offered for the data transmission and control and monitoring of wind turbines, we can adopt the following solution that encompasses the advantages of the above solutions:

The industrial solution that will apply uses a PLC

(FLOWTEL) with a management software (XFLOW) to secure and maximize the wind farms [13]-[17]:

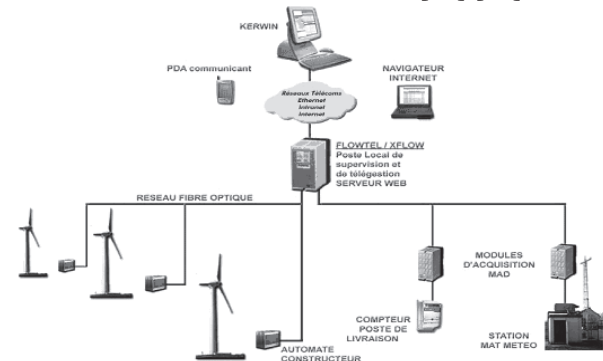


Figure 7. Structure of the remote management solution that will be applied to the wind farm in Tangier.

6. Conclusion

Wind energy is a major contribution towards energy independence of Morocco. So, the present works consists in feasibility study of harnessing wind energy for turbine installation in Tangier's site in northern Morocco. In the first of the study, we evaluated:

- The wind speed potential through the treatment of climatologically data of a station, using software ALWIN for dimensions of wind parks.
- The prediction of electrical energy produced in the site basically on the choice of wind energy converter.
- The determination of the direction of the wind energy converter in the site.

Indeed, we're studied the integration of a wind parks power production in the electric national network that have no negative effect on the electric grid. However, it empowers the electric national network capacity. Eventually, we're studied and suggested some solutions to apply remote monitoring of the wind farm to optimize our exploitation of the power production.

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